High-Voltage, High-Power, Solid-State Remote Power Controllers for Aerospace Applications

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NASA

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Summary

Two general types of remote power controller (RPC) that combine the functions of a circuit breaker and a switch were developed for use in direct-current (dc) aerospace systems. Power-switching devices used in these designs are the relatively new gate-turnoff thyristor (GTO) and power metal-oxide-semiconductor field-effect transistors (MOSFET's). The various RPC's can switch dc voltages to 1200 V and currents to 100 A. Seven different units were constructed and subjected to comprehensive laboratory and thermal vacuum testing. Two of these were dual units that switch both positive and negative voltages simultaneously. The RPC's using MOSFET's have slow turnon and turnoff times to limit voltage spiking from high di/dt. The GTO's have much faster transition times. All RPC's have programmable overload tripout and microsecond tripout for large overloads. The basic circuits developed can be used to build switchgear limited only by the ratings of the switching device used.

Introduction

Power demands of commercial, scientific, and military spacecraft are rapidly increasing. The present conceptual design of the space station power system projects a demand for 75 kW. It is estimated that spacecraft power systems may increase in capacity to 250 kW by the year 2000 (refs. 1 and 2). With higher power systems the distribution voltage will be increased to optimize efficiency and meet the system specific mass requirements. Some far-term applications may even require distribution voltages as high as 1000 V dc. New concepts such as high-voltage solar array switching (ref. 3) will also require higher voltage control systems. Although we will not see 1000-V dc systems for some time, the next generation of power systems will probably have distribution voltages between 100 and 300 V (refs. 4 and 5) and quite possibly at 270 V dc. Considerable development has already been done in this voltage range (refs. 6 to 11).

A critical need for high-voltage distribution systems is intelligent high-voltage, high-power remote power controllers (RPC's). NASA has funded several development programs to provide this technology. One of the early applications of remote power controllers developed by NASA was the 28-V dc distribution system of the space shuttle (ref. 12). Each shuttle uses about 600 of these units. The technology of high-voltage (>100 V) power controllers has been under development since the early 1970's and has resulted in the design of several units (refs. 13 to 15). The requirement is to switch high voltage at high power—and to do this in microseconds. One consequence of this requirement is that all-solid-state switching is mandatory. It not only eliminates the most important wearout mechanism, contact erosion due to arcing, but also permits operating fast enough to protect modern circuits.

Until recently, development of dc RPC's has progressed slowly, partly because no pressing need existed. NASA has funded several programs to develop high-voltage, high-power power controllers. Two of the initial high-voltage units used silicon-controlled rectifiers (SCR's) as the switching element (refs. 10 and 16). A major limitation of these units for space applications is their size and weight. SCR power controllers used in dc systems require a circuit to turn off or commutate the SCR. That circuit must provide considerable power at turnoff and is consequently large and heavy. Considerable work has also been done using SCR's in alternating-current (ac) power distribution systems (refs. 17 and 18), and at least one concept uses transistors in a Darlington circuit configuration (ref. 19). Alternating-current switching is not treated in this report.

Recently developed bipolar transistors, which may overcome the disadvantages of SCR power controllers, have been investigated. Because single bipolar transistors cannot handle the desired power levels, sophisticated bipolar designs with many transistors in parallel (ref. 9) are necessary. This is a viable approach for medium-power applications particularly when hybrid circuit fabrication techniques are used. The RPC's currently used on the space shuttle are of this type.

Recently developed high-power bipolar transistors (Westinghouse D60T and D7ST, sponsored by NASA Lewis) can switch significantly higher power but still have some limitations. Turning them on requires considerable base drive because of their low current gain. They are also physically large and heavy. A large spring is used within the device to provide good thermal contact between the unbonded semiconductor chip and its case. Because of this, the device cannot be mounted on a hybrid circuit to minimize weight. Metal-oxide-

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semiconducter field-effect transistors (MOSFET's) and gate-turnoff thyristors (GTO's) have a number of advantages over bipolar transistors and SCR's in switching applications. MOSFET's in particular can be used for many medium-power applications and can be paralleled for higher power. GTO's can handle both high current and high voltage in single devices.

This report describes the development of high-power, high-voltage dc remote power controllers that use MOSFET and GTO devices. These RPC's are more than just remotely controllable switches in that they also function as circuit breakers. They have programmable trip characteristics that allow matching to various load requirements. Trip characteristics are designed to protect both the load and the RPC. These devices are suitable for a wide variety of applications and help open the door to realizing high-voltage, high-power dc distribution systems in space.

All work was done in-house by the author with the technical assistance of Robert Gott, who contributed several valuable ideas as well as doing most of the construction and testing of the RPC's to be described.

RPC Design Features and Requirements

All of the switchgear described in this report was developed as part of an in-house program to extend the technology of solid-state switching. Each of the RPC's performs the combined functions of a remotely controlled switch and a circuit breaker. They are fully solid state and use no mechanical contacts. Two of the units were developed to control high-voltage power for electrostatic ion engines. Basic specifications are given in table I. These specifications generally cover all RPC's of both MOSFET and GTO types. Deviations and more detailed specifications are presented with each specific design.

Rated operating voltages and currents were chosen to cover a range that should include most near-term applications. Specific RPC ratings were determined by the switching devices available at the time. Overcurrent protection is provided in two steps. For moderate overcurrents a "slow trip" circuit integrates the overcurrent signal to give an approximate I^2t (current squared times time) response, similar to that of a thermal circuit breaker. When the overcurrent reaches a threshold, typically two to three times the switch rating, often referred to as two or three "per unit," a "fast trip" circuit takes over and the RPC opens in a few microseconds.

Besides meeting the basic switching requirements, the most important feature for RPC design is that the unit be self-protecting for all normal and fault conditions that may be encountered, including short circuits. The short-circuit requirement is the most difficult to meet. If the RPC is turned on into a short circuit, the current will rise at a rate determined by the source impedance and any line inductance, unless limited by the RPC. If the switch element is a transistor or similar device with controllable turnon time, a soft turnon can be programmed. An overload or short circuit can then be sensed before the current reaches destructive values, and the switch can be turned off. When GTO's or SCR's are used, there is no control of the turnon time and the trip circuit must be fast enough to act before destructive currents are reached. If there is little inductance in the power circuit, this may be difficult. For the GTO RPC's described herein, the overcurrent trip circuit was made to operate as fast as possible. The RPC is then restricted to use in

TABLE I.—GENERAL RPC SPECIFICATIONS

Rated operating voltage, V dc	150 to 1200
Rated output current, A dc	≤100
Control power source, V dc	28 ± 7
Low control power protection	Shutdown for <21 V
Control voltage	TTL compatible
On-and-off control signals	Logic "O" (low)
Overload trip indication	TTL signal—high after trip
Overload reset	Resets on "off" command
Turnon delay, ms	100 max.
Rise and fall times, μs	100 desired
Voltage drop at rated load, V dc	2.0 max.
Power dissipation off, W	0.5 max.
Efficiency at 20 to 100 percent of rated load current, percent	≥99 (including control power)
Overload tripout	Proportional to I^2t
Fault response time, μs	3
Fault trip level	2 to 3 per unit
Operating temperature, °C	−50 to 75
RPC protection	RPC must be self-protecting under all conditions including maximum fault current

systems that either have inherent current limiting or have sufficient inductance to somewhat limit the rate of current rise.

Another consideration is the type of load to be controlled. Resistive and inductive loads pose little problem. Many loads may have capacitive input filters, which are not as easily controlled. Large surge currents that may exceed the overcurrent limit occur at turnon. Here again, the "soft" turnon feature solves the problem. It charges the capacitive load slowly, without exceeding the tripout current. At turnoff a slow switching time is also desirable, particularly for inductive loads. When the load circuit is opened, an inductive surge voltage proportional to the inductance and rate of change of current is produced. The slow turnoff, practical with the MOSFET RPC, minimizes this surge. The GTO RPC's turnoff is not controllable and is quite fast. GTO's are readily available with voltage ratings above 1 kV, so they are more tolerant of voltage surges. It is possible to clamp voltage spikes with zener diode surge-voltage suppressors, and this has been done in most of the designs.

It was not the intention of this program to build flight hardware as an end product. As it turned out, the RPC's that were built had to be carefully packaged to meet circuit requirements. The resulting units were subjected to and passed thermal vacuum tests. No shock or vibration testing was done although the construction practices followed should be adequate to meet these. All units could probably be qualified as they stand. If they were properly redesigned mechanically, some saving of weight and volume could be realized. If they were redesigned in hybrid form, the weight and volume savings would be appreciable, possibly achieving reduction factors of 2 to 5.
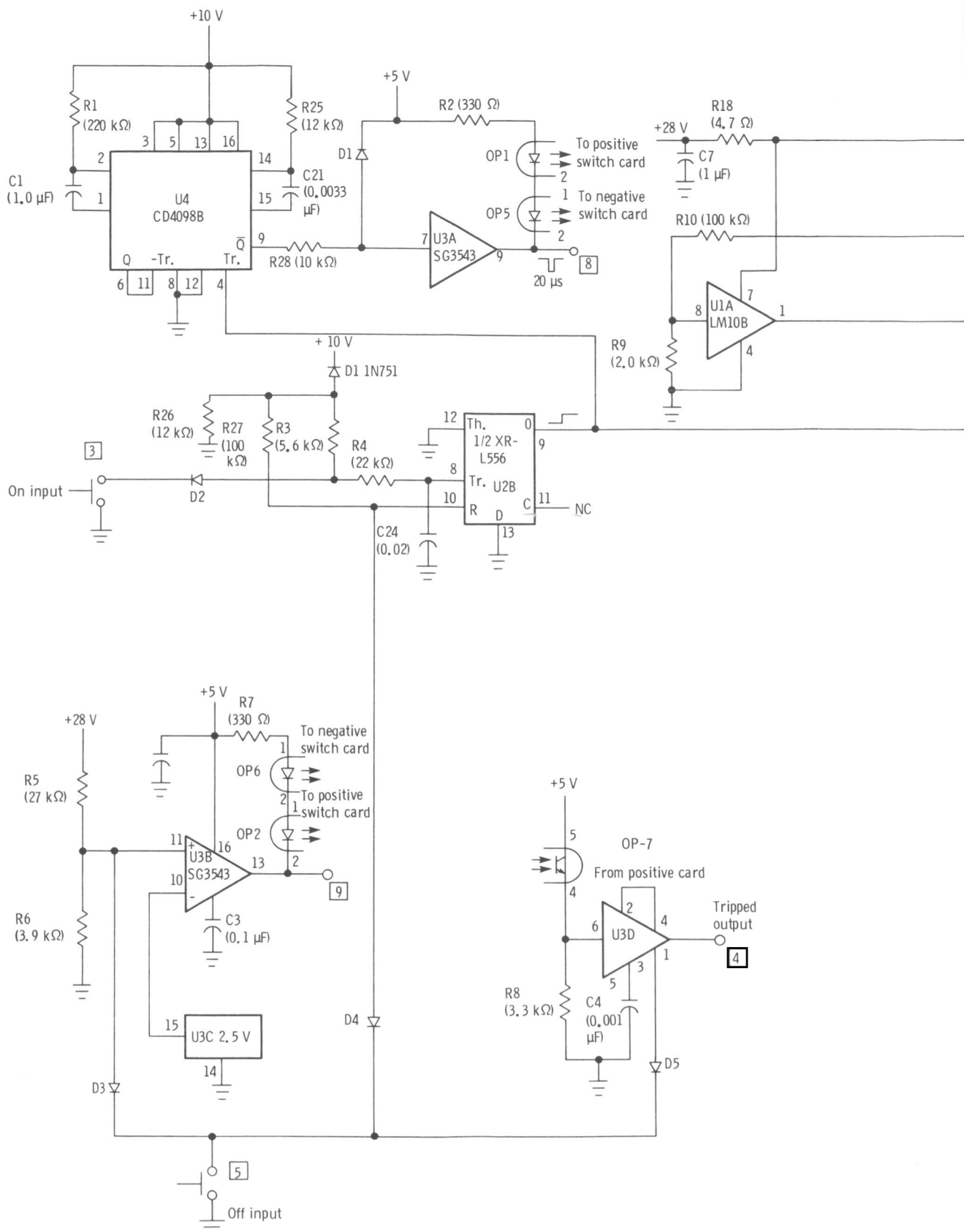
Design of Common Elements

Control requirements for all of the RPC's were standardized so that a single basic circuit could be used for all seven devices. Separate on-and-off control lines plus a latched output indicating overload tripout were provided. All logic is transistor-transistor logic (TTL) compatible and the inputs can also accept a switch closure to ground. Of necessity, the input and output must be referenced to the common control power ground, which is also the high-voltage power return for most of the RPC's. The two units developed for ion engine power control have a common isolated power return that may be biased to 100 V off ground. This circuit does not require any changes in the control circuitry. It also interfaces the ground-referenced on-and-off commands to optical isolators on the individual switch cards, provides turnon and turnoff delays, and has an output that indicates when the RPC has been tripped off.

One of the design decisions that had to be made was whether the current sensing and trip circuits should be referenced to ground or should "float" on the high-voltage power line. If they were ground referenced, isolated power supplies would not be required. In fact, the control circuits could then be operated directly from the 28-V dc power, and most of the power-conditioning requirements could be eliminated. For this to be practical requires some means of sensing current that produces a ground-referenced output. Several techniques that use magnetic coupling or Hall sensors are available to do this, but none have the fractional microsecond response times necessary for this application. The obvious alternative of sensing current in the ground return was also rejected because this precludes using the vehicle structure as the return. It might also be possible to transfer a current measurement from a floating sensor to ground via an optical isolator. Most circuits that would do this require dc power at the floating sensor. The optical isolator also introduces a small but not negligible time delay.

After the drawbacks of these current-sensing approaches had been considered, a resistive shunt with floating sense electronics was selected. All of the switching and control circuitry "floats" above ground and is referenced to the high-voltage switched output. Isolated power is provided from a high-frequency inverter through a small transformer. As many as four different isolated voltages are supplied for the ion engine power controllers. Signal transfer to and from the "floating" switch circuit is via optical isolators.

Control Circuit and Power Supply

A typical control circuit and power supply is shown in figure 1. The power supply provides regulated power for the control circuit and also isolated power for two "floating" switch drive circuits. In this case, ± 12, $+10$, and -5 V are supplied with 1200-V isolation to ground and 2400-V isolation between the two power supply outputs. There are also two ground-referenced power levels. One is a 10-V level, regulated down from the 28-V control power input, which is energized whenever control power is applied. The second is 5-V power from the switching converter. There are two possible modes of operation. Both sections can be wired to turn on whenever control power is applied, or the 5-V section can be controlled by the on/off inputs. In the latter mode the power consumption in the off-state is reduced to about 100 mW. This is the power for the turnon circuit, which must be continuously energized to accept an on-command. The turnon circuit consists of U1, U2, and U4, which are powered directly from the 28-V line. The remaining control circuit, U3, is powered from the 5-V converter output.

Figure 1.—Schematic diagram of control circuit and

C8 (12 μF, 50 V)

R11 (24 kΩ)

R12 (24 kΩ)

+10 V

R17 (3.3 kΩ)

R20 (6.8 Ω)   +5 V

T1   D8

8T   C10 (2.2 μF)   C11 (68 μF)

U1B   6   3

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C21 (0.1 μF)

D9   R21 (10 Ω)   +12 V

18T   C12 (2.2 μF)   C13 (1.0 μF)

R16 (1 kΩ)   R13 (100 kΩ)

C6 (0.001 μF)

C9 (270 pF)

40T

R19 (100 Ω)

To positive board (+1200-V isolation)

C14 (2.2 μF)   C15 (1.0 μF)

R14 (33 kΩ)   4   14

1   R   C

D   3

18T   R22 (10 Ω)   -12 V

D10

D6   R15 (100 kΩ)

1/2 XR-L556

U2A

6   Tr.

2   Th.   0   5

D11   R23 (10 Ω)   +10 V

15T   C16 (2.2 μF)   C17 (1.0 μF)

C20 (0.01 μF)   D7

C5 (330 pF)   7

Q1 1RF611

To negative board (-1200-V isolation)

C18 (2.2 μF)   C19 (1.0 μF)

9T   R24 (10 Ω)   - 5 V

D12

power supply. All diodes are 1N916; all optoisolators are MCT2E.

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A ground applied to pin 3 turns the RPC on. High-frequency noise is limited by R4 and C24, after which the input command trips U2B, which is used as a latch. Setting the latch gives a high output on pin 9 and enables U2A, which is a pulse-width-modulated multivibrator that directly drives the single-power MOSFET, Q1, in a flyback converter. A flyback converter was chosen for simplicity and fewest components. One of the converter outputs is 5 V, which powers the remainder of the ground-referenced control circuits. Amplifier U1B regulates the 5-V supply by comparing it with 5 V divided down from the 10-V supply by R11 and R12. The error signal from U1B pulse-width modulates U2A, which in turn controls the on-time of the switching transistor Q1. During steady-state operation the converter frequency is approximately 30 kHz and the on-time is about 4 μs.

After an on-command is received, a delay must be provided before the output switch is commanded on, to allow the power supplies to come up to full voltage. This is especially important should the RPC be turned on into a short circuit. Sufficient energy must be available to the turnoff circuitry to turn off the switching device very rapidly if a failure is to be avoided. Dual one-shot multivibrator U4 supplies the required delay. Its first section is triggered on by the start command. When this section times out in about 100 ms, it triggers the second one-shot multivibrator, which outputs a 20-μs pulse. Comparator U3A amplifies this pulse and can drive the on-optoisolators of two separate switches.

Turnoff is handled similarly. Comparator U3B is an undervoltage detector with a built-in 2.5-V reference, U3C. Its input is normally biased above 2.5 V by R5 and R6 from the 28-V line. Should the line voltage drop below 21 V, the output of U3B will go low, turning on the optoisolators OP2 and OP6 and turning off the respective switches. In this way the RPC will turn off if control power is lost. It is not possible to lose power in such a way that the ability to turn off the RPC is lost. Nor can it get into a high-power dissipation state. Normal turnoff, a low level at connector pin 5, produces the same effect as loss of control power and uses the same circuit (U3B). U3B consists of two comparators. The first one outputs a 250-μA current when tripped. This charges capacitor C3. When the voltage on C3 reaches 2.5 V, the second comparator trips and drives the output low. Capacitor C3 sets a delay that prevents false tripouts from noise pulses on either the control or 28-V lines. The minimum input pulse width for turnoff is 320 μs. A low level at the off-input does two other things. It resets the latch (U2B) that controls the power converter, turning it off. It also resets the tripout indicator U3D if it had been set previously.

Tripout due to overload is handled by the power-switching circuit and is described under MOSFET RPC Circuit Description. If there are two power switches, they are interlocked via optoisolators so that if one trips out, the other one will do likewise. This is an important feature for RPC's intended for ion engines. They could be damaged if one of the high voltages was lost and the other remained on. When a tripout does occur, a signal is coupled via OP7 to the ground-referenced control card and sets latch U3D. It provides the output indicating tripout, which remains set until the RPC is commanded off.

RPC Construction

Very early in the program it became clear that the original idea of developing and testing breadboard circuitry would not work. The requirement of sensing low voltage levels across the current shunt during the switching process, with circuits isolated from ground, made breadboard operation impractical. Pickup and circuit coupling could not be adequately controlled. This was particularly true for the GTO RPC's because of their fast switching times. It was therefore decided to build all test RPC's in a semiflight configuration. Little effort was spent in minimizing package size and weight, and all units were made very similar. One standard printed-circuit (PC) card size was adopted, and the only size variations were to accommodate different power-switching devices. An added advantage of this form of construction is that the units could be thermal vacuum tested. Although the packaging was very adequate for development purposes, improvements are possible. The greatest improvement would come from a hybrid circuit approach, using ceramic substrates and chip semiconductors for the power switches as well as the control electronics. This is discussed in more detail in the section Hybrid Packaging Option. Figure 2 shows typical RPC construction for a two-circuit (type A) unit. Figure 3 shows three complete RPC's: types B, P, and F.

MOSFET Remote Power Controllers

Advantages and Disadvantages of MOSFET Devices

The availability of high-power MOSFET's with breakdown voltages of several hundred volts has made possible RPC's with significant improvements over those using bipolar transistors. One of the prime advantages of power MOSFET's is the very low drive power required in both the on and off states. Because the input impedance is almost purely capacitive, little power is required except during turnon and turnoff. Input capacitance, as high as 5000 pF for a single transistor, causes high surge currents during transition. However, this power can be stored in a capacitor to provide the surge current. Additionally the low drive power gives higher efficiency than a bipolar switch. Other advantages include fast and controllable switching times, ease of paralleling multiple devices, and the potential for hybrid packaging.
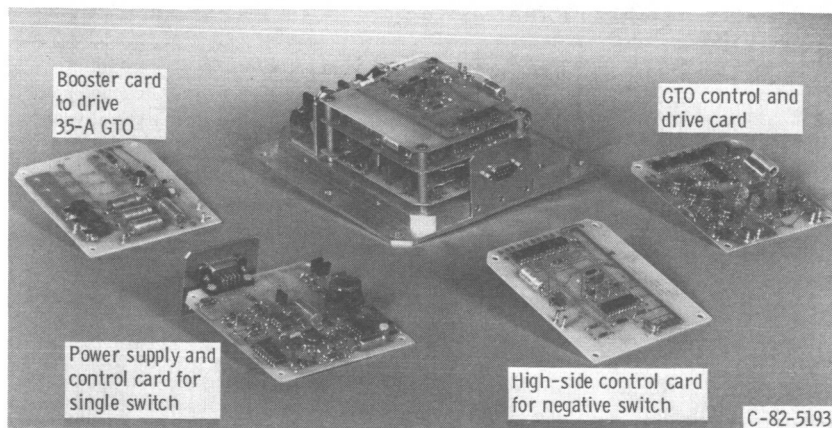
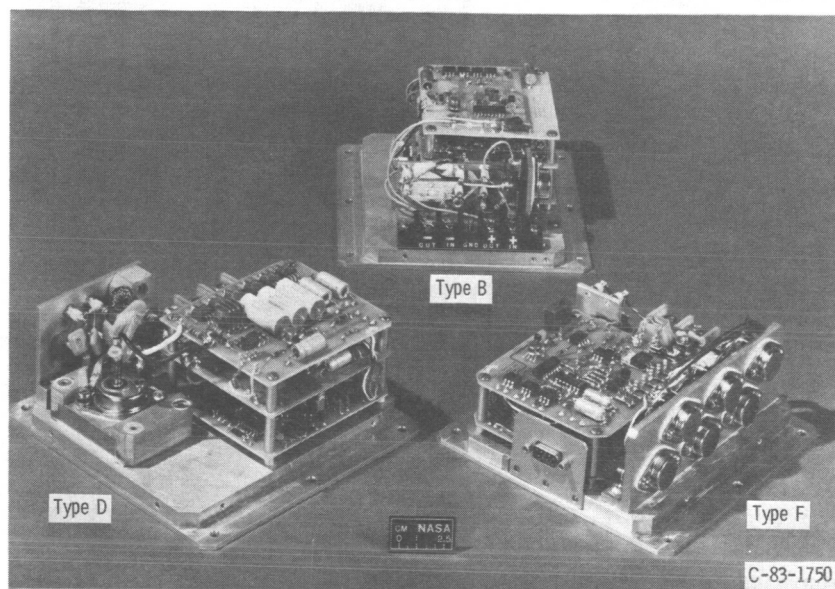Figure 2.—Typical circuit cards and two-circuit remote power controller (type A).

Figure 3.—Remote power controller types B, D, and F.

As part of this study, a survey of suitable power MOSFET's was made. Some of the better transistors are listed in table II. These include the devices used in the RPC's to be described.

MOSFET's also have a few disadvantages. The positive temperature coefficient of the drain-to-source resistance $R_{DS(on)}$ limits current rating or junction temperature. If a given maximum forward-voltage drop is to be maintained, operating current must be traded against junction temperature. In a typical RPC the forward-voltage drop will double for a junction temperature change from 25 to 130 °C. For a design that must operate at a 130 °C junction temperature, this means that the data sheet value of $R_{DS(on)}$ must be doubled. The other disadvantage of MOSFET's is cost. Currently the cost of switching power is higher with MOSFET's than with bipolar transistors. Some price reductions have already been made, and it is expected

that continued competition and expanding demand will bring prices to competitive levels in a few years. A minor disadvantage is the limited current that can be carried by a single MOSFET. When it is required to keep forward-voltage drop between 1 and 2 V, maximum current must be limited to much less than advertised values. As a result many transistors in parallel are required to switch high currents.

Other Devices

In addition to MOSFET's, two similar types of device were evaluated. They are the Supertex SUPERFET SNO140N1, and the General Electric insulated gate transistor (IGT) D94FQ4. Both appeared to have advantages on the basis of published current ratings. The main feature is a higher current density per unit area and therefore a favorable performance/cost tradeoff. A

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TABLE II.—MOSFET TRANSISTORS CURRENTLY AVAILABLE

Device	Drain-to-source voltage, V_{DS}, V	Drain-to-source resistance, $R_{DS(on)}$, Ω	Case temperature, °C		Input capacitance (typical values) C_{iss}, pF
			25	100	
			Drain current, I_D, A		
MTP1N100	1000	10	1	0.4	1200
BUZ58A	1000	2.6	3.7	2.4	3500
BUZ88A	800	1.5	5.0	3.2	3500
IRF450	500	.4	12	7.75	2000
VNP006A	500	.3	20	14	4500
IRF350	400	.3	14	9	2000
VNM005A	400	.2	25	17	4500
IRF250	200	.085	30	19	2000
VNJ004A	200	.060	45	32	4600
MTE60N20	200	.048	55	----	7000
VNE003A	100	.035	60	45	4600

closer evaluation (fig. 4) shows that both have a higher saturation voltage at low currents than the IRC IRF350 MOSFET. At about 6 A the SNO140N1 has the same drop as the IRF350, and the D94FQ4 is considerably higher. The G.E. and Supertex devices are slower than the IRF350, and as would be expected for devices with a smaller chip size, they have considerably smaller safe operating areas. These are important considerations during overload trip. It was concluded that, of the devices currently available, the pure MOSFET's were best for dc systems.

MOSFET RPC Design Philosophy

The forward-voltage drop is the feature most often used in comparing the performance of RPC's. Bipolar transistors or thyristors have an inherent saturation voltage that fixes the minimum forward-voltage drop. This is not true for MOSFET's. In the on-condition MOSFET's appear predominantly resistive, so that an arbitrarily low voltage drop can be attained by paralleling enough transistors. An example will illustrate the design. Assume the following switch specifications are required:

Forward current, A dc 100
Voltage rating, V dc 48
Maximum case temperature, °C 85
Maximum forward-voltage drop, V 1.0

Start by selecting the Siliconix VNE003A transistor. Assuming a maximum junction temperature of 100 °C leads to an increase in $R_{DS(on)}$ of 1.7 times the 25 °C value of 0.035 Ω. Thus the maximum $R_{DS(on)}$ is 0.059 Ω. For a forward-voltage drop of 1.0 V this transistor will therefore carry 16.8 A at 100 °C. Six will be needed for a 100-A switch. This is a very conservative operating point for this transistor, which is rated at 45 A at 100 °C. Maintaining the maximum junction temperature requires that the thermal drop from the junction to the case be held to 15 deg C, which corresponds to a thermal
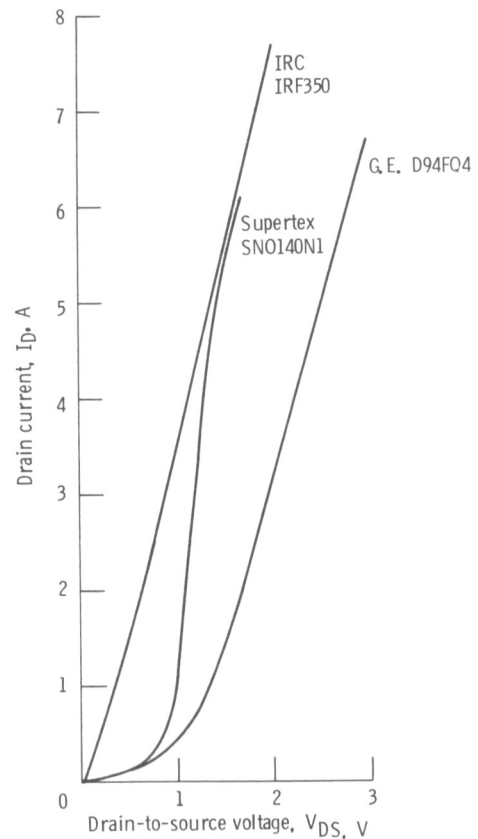

Figure 4.—Saturation characteristics of alternative MOSFET devices.

resistance of just under 1 deg C/W, a realistic value. At 25 °C the forward-voltage drop will only be 0.59 V. Thus the limiting factor of this design is thermal.

One further limitation must be considered (i.e., current sharing among the transistors). It is often assumed that the positive temperature coefficient of $R_{DS(on)}$ will assist in balancing the steady-state current among the parallel MOSFET's. This is only true above a crossover current

where the temperature coefficient changes sign, as can be seen clearly in the transfer characteristic for the IRF320 MOSFET at a drain current of 3 A (fig. 5). Thermal limitations usually dictate operation below this critical current. A detailed discussion of this problem can be found in reference 20. Practical solutions include selecting transistors so that they share current fairly closely at the gate drive used and further thermal derating. Reference 20 also discusses active thermal control, which could be used in hybrid construction to maximize ratings.

No difficulty in matching transistors from small lots was experienced in the construction of eight units. The procedure used was to measure the gate-to-source threshold voltage V_{GS} required to conduct a given current. The current at a V_{GS} of 2 V was also measured. Transistors were then selected to have approximately the same threshold voltage. In addition, transistors with abnormally high or low current at 2-V V_{GS} were not used. For one 35-A RPC using the VNM005A MOSFET's, the spread in threshold voltages was 0.7 V.

Current sharing during turnon and turnoff should also be reasonably matched among paralleled MOSFET's. This is usually not a problem, particularly if they have been statically matched (ref. 21). For operation at conservative currents, as in the preceding example, a 25-percent unbalance during transition poses no problem. Transistors that were statically matched with this procedure were also acceptably matched during turnon and turnoff. Oscilloscope traces showing the individual transistor currents for one unit are included in the testing section.

Turnon and turnoff of purely resistive loads causes no problems, but practical loads are seldom resistive.
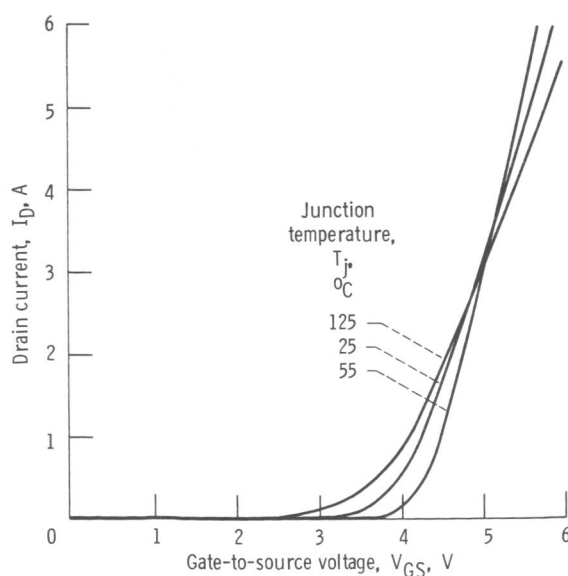
Figure 5.—Transfer characteristics of IRF320 MOSFET. 80-μs pulse test; $V_{DS} > I_{D(on)} \times R_{DS(on),max}$.

Capacitive loads, such as the input filters in equipment being controlled, draw large surge currents when turned on rapidly. Even small capacitors will draw more than the three-per-unit trip current. In operation the RPC appears not to turn on. Actually it turns on and off in a few microseconds. One means of switching capacitive loads is to incorporate current limiting in the RPC. Although this has been done at low currents (ref. 9), it complicates the design and requires considerable thermal dissipation. A more practical solution is to use a slow turnon, which has much the same effect as current limiting. For the MOSFET RPC's this was implemented by driving the gates with a ramp voltage. The same thing was done for turnoff. This limited the rate of change of current in inductive loads and prevented the generation of large spike voltages that would damage the MOSFET's. An example will illustrate the latter feature. If a switch carrying 100 A in a circuit with an inductance of 20 μH is opened in 5 μs, a transient voltage of 400 V will be generated ($L \, di/dt$). As can be seen from table II, MOSFET's with high enough voltage ratings to handle this magnitude of surge voltage on top of the normal line rating are not available. Slowing the turnoff time to 50 μs decreased the spike voltage generated to 40 V, which can be more readily handled. A minimum turnon and turnoff time of 50 μs was selected. Longer times are used for some RPC's. If long switching times are used, it may be necessary to further derate transistor current to stay within the safe operating area of the transistors.

For moderate overcurrents it is desirable to have an RPC mimic the action of a conventional circuit breaker. These units use an enhanced RC type of integrator that approximates an I^2t response. Overloads of 115 percent will take several seconds to cause a tripout. As the overload increases, the trip time decreases. A three-times-rated current trip time is 50 ms. Up to this level the slow-turnoff feature is still operative. Above this level a second, fast-trip circuit is enabled. It must function very quickly to protect against large overloads and particularly against short circuits.

When an RPC turns on into a short circuit, the current will rise at a rate controlled by the source and line inductances. These are usually too small to sufficiently limit the rate of current rise. It is therefore necessary to sense large overcurrents quickly and turn off the RPC before destructive current levels are reached. The soft turnon possible with MOSFET's makes this easy. For a 300-V, 35-A RPC, di/dt into a short circuit is limited to 0.3 A/μs by driving the MOSFET gates with a ramp voltage. In this case it would also be possible to use a slow shutdown to limit voltage spiking. The added circuit complexity and the requirement to protect against short circuits when the RPC is fully on make this less practical.

A "hard" short at full load with the RPC fully on causes the current to rise at 30 A/μs. If the fast-trip circuit responds in 1 to 2 μs, the current will not reach

destructive levels. This effectively limits peak let-through current but creates a transient voltage problem because of the energy stored in line inductance. With a 15-A/μs di/dt a voltage transient of several hundred volts can be generated. The transistors can be protected by using zener surge-voltage suppressors. An alternative approach could be used for a specific application where the source and line inductances are known, that is, to optimize the fault turnoff time for the particular inductance. Since a larger inductance would limit the rate of rise of short-circuit current, more time would be available for turnoff before a destructive current was reached. Using a longer time would limit the overvoltage produced.

Specifications of MOSFET RPC's Built

MOSFET switches were used in four of the RPC's. Three models—E, F, and G—were high-power units. The fourth model, A, used a MOSFET to switch a low-current, high negative voltage used for the accelerator grid of an electrostatic ion thruster. The high-power, high positive voltage for the screen grid was switched with a GTO switch. Specifications for the MOSFET RPC's are given in table III. Note that two different MOSFET's were used for type E RPC's. The VNM005A's have a lower forward-voltage drop but are more prone to failure than the IRF350's. Both are referred to in various sections of this report.

MOSFET RPC Circuit Description

One of the design goals was circuit simplicity without sacrificing performance. Keeping the parts count low will not only minimize size and weight, but will also improve reliability. All of the RPC's use the same power supply and control electronics. The three high-power MOSFET RPC's have nearly identical circuits differing by only a few component values. Each has six TO–3-cased

TABLE III.—MOSFET RPC SPECIFICATIONS

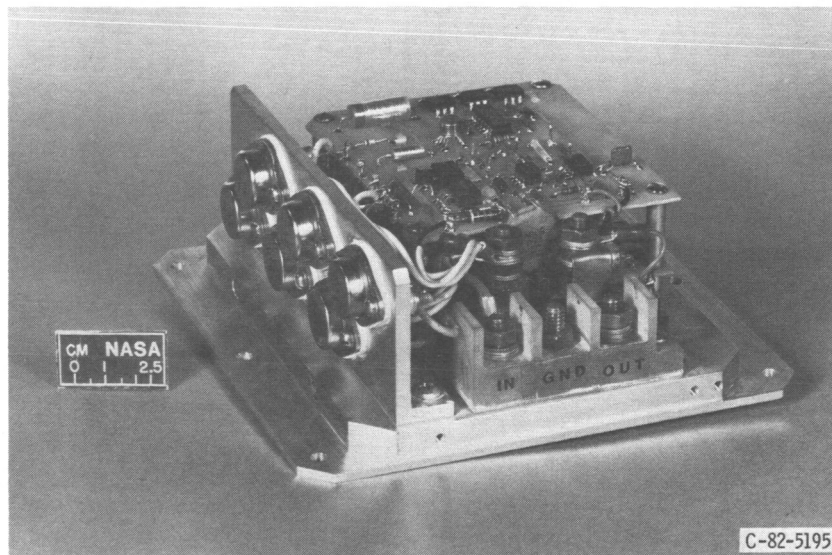|	RPC model			
	A	E	F	G
Operating voltage, V	−750	300	150	800
Full load current, A	0.04	35	100	10
Switching time (on/off), μs	5	50	250	100
Fault turnoff time, μs	120	2	3	3
Switch-on drop at full load, V	0.25	1.7	2.0	3.8
Efficiency, percent	99.9	99.4	98.7	99.5
Device	MTP1N100	IRF350 or VNM005A	IRF250	BUZ54A

MOSFET's in parallel for the power switch. The number was chosen to be large enough to demonstrate the practicality of paralleling MOSFET's, to give a reasonable current rating, and to package easily. There is no reason that more or fewer devices could not be used with this design, as long as the turnon/off drive capability of the circuit is not exceeded. A typical RPC is shown in figure 6.

The type E (35-A, 300-V) RPC, shown schematically in figure 7, is used here to describe circuit operation. Turnon and turnoff signals are coupled from the power supply and control card to the floating switch card by optoisolators OP1 and OP3. When OP1 is pulsed, after the turnon time delay, it pulls up the clock input, pin 3, of flip-flop U4, which sets the Q output low. This flip-flop controls the on-function of the switch and is set whenever the switch is on. It would be possible to drive one or more MOSFET's directly from the output of U4 or to drive a number via a high-power buffer. The difficulty is that the resultant turnon and turnoff times are too short and cannot be extended sufficiently by adding series resistance in the gate leads. U6 is an integrator that generates a ramp voltage to control turnon and turnoff. Its noninverting input is set at one-half the 12-V supply voltage by the voltage dividers R19 and R20. When the input from U4 goes to zero, the integrator output starts at zero potential and ramps up to very nearly supply voltage. The slope of the ramp is set by C10 to give the desired turnon time. Because the turnon threshold of the IRF350 MOSFET's is about 3 V, no output change occurs until the ramp reaches this level. Turnon occurs during the short interval before saturation is reached. Only a small portion of the range corresponds to the actual turnon time. Using an operational amplifier, U6, as the driver for the MOSFET's also provides a low drive impedance.

Small resistors, R21 to R26, are added in series with the IRF350 gates to suppress oscillations. Each gate is also protected from excess voltage by a zener diode (1N5242B), which limits the maximum voltage to 12 V. A snubber is used across the switching transistors to slow down the rate of voltage rise at turnoff. In addition, four high-power zener diode surge-voltage suppressors protect against transient voltages, and a diode clamp to ground prevents reverse voltage. All leads in the power circuit are kept as short as possible and are positioned for minimum coupling to the control circuit. Load current I_L is sensed by current shunt R10. It is sized for 87 mV at a full load current of 35 A. When a turnoff command is processed by the ground-referenced control circuit and power supply card, it turns on optoisolator OP3. This pulls pin 1 of NAND gate U5A low and its output goes high. This signal is inverted by U5B and clears flip-flop U4A. The RC network at the clear pin (1) of U4A is not used to delay the turnoff signal. Capacitor C13 is necessary to ensure that U4A is initialized in the off-state when power

Figure 6.—MOSFET remote power controller.

is first applied. The turnoff command thus resets U4A, which outputs a high level at $\overline{Q}$, pin 6. Operational amplifier U6 integrates this signal and generates a negative-going ramp that turns off the power switches. With the noninverting input of U6 held at one-half of supply voltage, the logic level input (0 or 12 V) is equivalent to a positive or negative 6-V signal. Turnon and turnoff ramps therefore have the same slope. If the reference voltage applied to pin 3 of U6 was at other than half supply voltage, the turnon and turnoff ramps would have different slopes and the relative switching times could be appropriately controlled.

Normal overcurrent tripping with an inverse time delay is controlled by dual operational amplifier U1. The first stage, U1A, is configured as an augmented integrator. When a resistor, R4, is added in series with the feedback capacitor, C1, of an integrator, the circuit produces a composite output. This output contains a component proportional to the input signal, which is added to a component proportional to the time integral of the input signal. Proper selection of the proportional and integral components approximates the desired I^2t trip characteristic. Resistors R1, R2, and R4 and capacitor C1 determine this trip characteristic. To prevent tripout below the rated current setting, a bias current is established by resistor R3 at the input to the integrator. A precision voltage reference, U2, ensures that this bias will remain stable with time and temperature changes. The bias is positive, and R3 is chosen so that it just matches the negative current signal from the shunt R10 at full load. Below full load a net positive current flows into the integrator input and drives the output negative. The U1 operational amplifier is, however, operated from a single 12-V supply, so the output cannot go below ground. It stays at ground potential during normal operation. When

a small overload occurs, the input bias is overcome and a negative input current flows into U1A. It is integrated and the output at pin 1 ramps positive. U1B is used as a comparator, with its reference input, pin 5, set at 5.0 V. When the signal from the integrator exceeds the 5.0-V reference, the output at pin 7 goes to ground. This has the same effect as an off-command from the optoisolator, as described previously. For large overloads the augmented integrator outputs an immediate step superimposed on the ramp, which trips the circuit off sooner than if it were controlled solely by a linear integrator. Tripout time decreases with increasing overload up to a three-per-unit current. At this current the tripout time is 50 ms and turnoff takes 50 μs.

An output is required from the RPC indicating when it has turned off because of an overload. The positive-going pulse at U5A, pin 3, in the overload trip circuit is used. It clocks U4B, which is used as a one-shot multivibrator. The Q output of the multivibrator drives optoisolator OP2, the output of which latches a circuit on the control and power supply card as described previously.

When the switch current exceeds the three-per-unit level, the fast-trip circuit is activated. Using the same voltage sensed across current shunt R10, it turns off the RPC in less than 2 μs. This voltage feeds a high-speed comparator, U3. Its input is biased from the precision 5-V reference via resistor R9 so that its threshold corresponds to the three-per-unit current level. When that level is reached at the input, its output goes positive. The output does two things: it directly drives the gate of Q1, turning it on, and shorts out the gate drive to the switch transistors, turning them off. Because of the very fast action of this circuit two feedback loops are necessary to ensure that the switch stays off. The first is positive feedback around U3. Its output signal is coupled
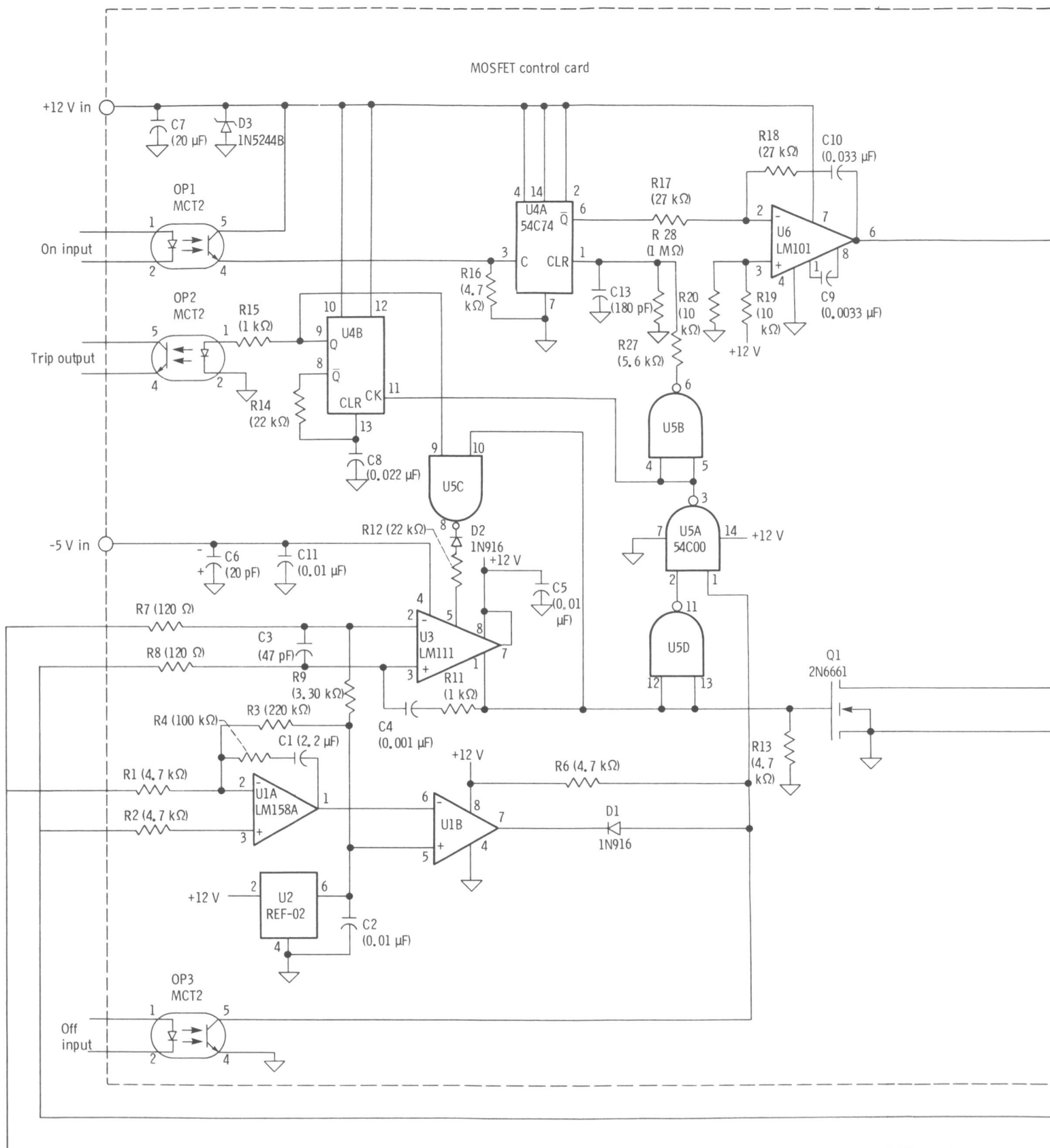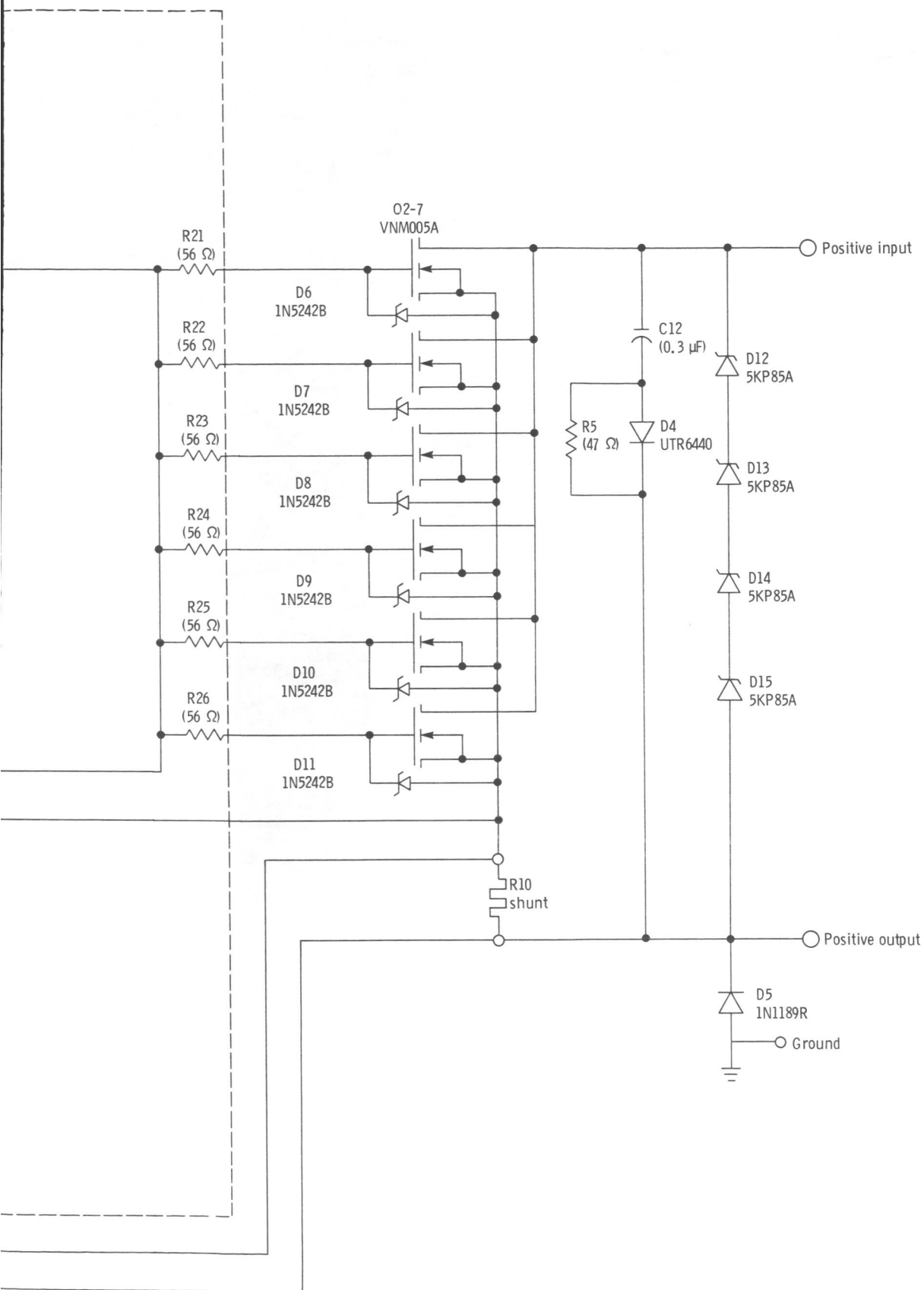
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Figure 7.—Schematic diagram of type E MOSFET remote

O2-7
VNM005A

R21
(56 Ω)

D6
1N5242B

R22
(56 Ω)

D7
1N5242B

R23
(56 Ω)

D8
1N5242B

R24
(56 Ω)

D9
1N5242B

R25
(56 Ω)

D10
1N5242B

R26
(56 Ω)

D11
1N5242B

C12
(0.3 μF)

R5
(47 Ω)

D4
UTR6440

Positive input

D12
5KP85A

D13
5KP85A

D14
5KP85A

D15
5KP85A

R10
shunt

Positive output

D5
1N1189R

Ground

power controller. Current, 35 A; voltage, 300 V dc.

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back to the noninverting input by R11 and C4 to ensure fast switching and to maintain the switched state for a few microseconds. Note that the input polarity is reversed from normal because the comparator is configured to provide a positive output at pin 1. A second positive feedback loop, through U5D, U5A, U4B, and U5C, also holds the output high for the duration of the one-shot pulse from U4B. At the same time, U5B clears control flip-flop U4A and removes the drive from U6. It takes operational amplifier U6 a number of milliseconds to fully turn off internally even though its output has been forcefully pulled to ground. While this is happening, the positive feedback loops ensure that Q1 stays on and maintains the switch off.

Experimental Evaluation of 35-A MOSFET RPC

Each RPC was subjected to thorough bench testing covering the full range of operating conditions. For the MOSFET units this included current sharing, static forward-voltage drop, normal turnon and turnoff, turnon and turnoff of a capacitive load, turnon and turnoff of an inductive load, slow overcurrent trip, fast overcurrent trip, turnon into a short circuit, short circuit from full load, and thermal vacuum tests. The test setup consisted of two Electronic Measurements Inc. 500-V, 60-A SCR type of power supplies and a resistive load bank. Each of the supplies could be operated independently, or connected in series or parallel via a patch panel. Interconnecting wiring was number 4 AWG welding cable. Some stability problems were encountered when running high-current tests. When an RPC cleared a heavy overload condition and switched off in a few microseconds, the power supply output voltage tended to overshoot until the control circuit could fully recover. To prevent this, an RC filter consisting of a 100-μF low-inductance capacitor in series with a 1-Ω resistor was added across the supply leads near the test RPC. This minimized voltage spikes that otherwise might have damaged the RPC's.

Current sharing.—Typical current sharing for a 35-A RPC using six parallel VNM005A MOSFET's is shown in figure 8. They were matched as previously described. All currents were measured with standard Tektronix current probes. The top six traces are the individual transistor currents and the bottom trace is the total current. Transistor 4 was the closest to average threshold and its current followed closely the total current. Transistors 1, 2, and 3 had somewhat lower threshold voltages. They therefore turned on sooner and conducted more current during the turnon transient. Likewise, transistors 5 and 6 had higher gate threshold voltages and were slower to turn on. They therefore conducted less than nominal current during turnon. Since the operating currents of each transistor were well under the maximum rated current, these deviations from ideal current sharing are
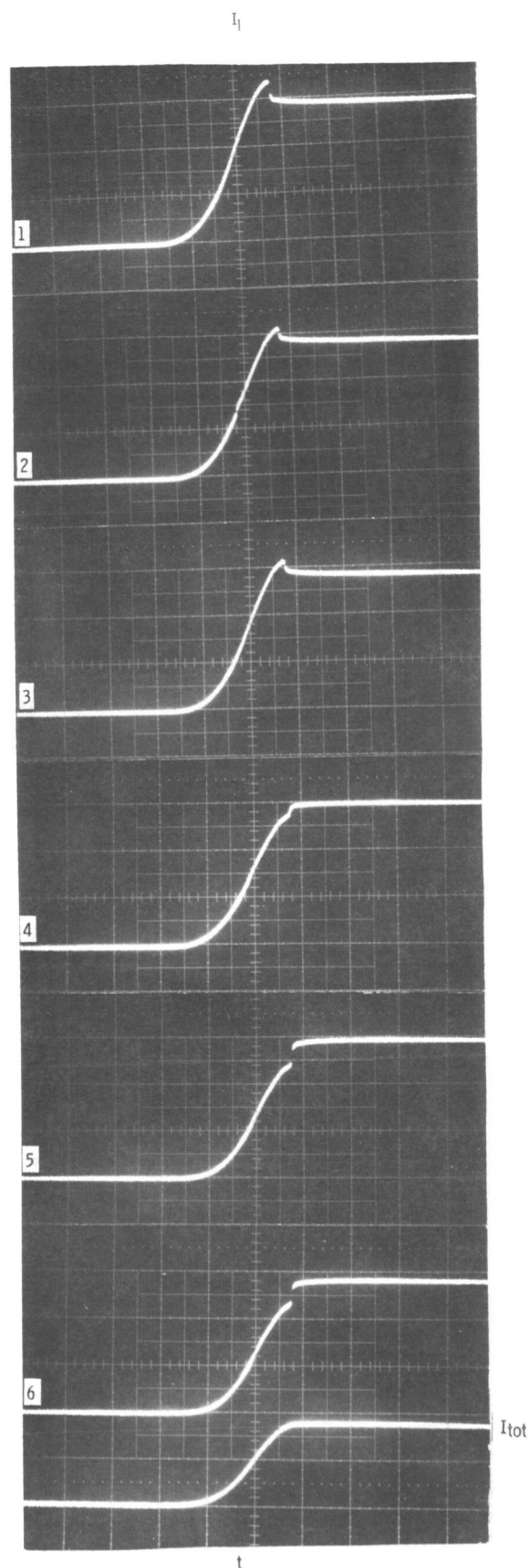

Figure 8.—Current sharing among six parallel VNM005A MOSFET's. Upper six traces are individual transistor currents at 2 A/division. Lowest trace is total current at 20 A/division. Sweep, 20 μs/division.

inconsequential. After the turnon transient the transistors shared current inversely proportional to their on-resistances. Measured values of saturation current for the six transistors in this switch ranged from 4.6 to 5.4 A at a gate voltage of 10 V and a saturation voltage of 1 V. These measurements were made with a Tektronix transistor curve tracer. They compare with the parts of the individual current traces shown in figure 8.

Static forward-voltage drop.—Static forward-voltage drop was also measured over the full range of currents at several temperatures. Figure 9 shows the values for RPC E2 at 25 and 100 °C. The forward-voltage drop corresponded closely to a constant ohmic resistance. Forward-voltage drop is the parameter that determines the current rating of the RPC. For aerospace applications where power loss is a major design factor, MOSFET's cannot be operated at their rated current. For instance, the VNM005A MOSFET is rated to continuously carry 17 A at a 100 °C case temperature. Under these conditions the saturation voltage would be 4.6 V. This is too
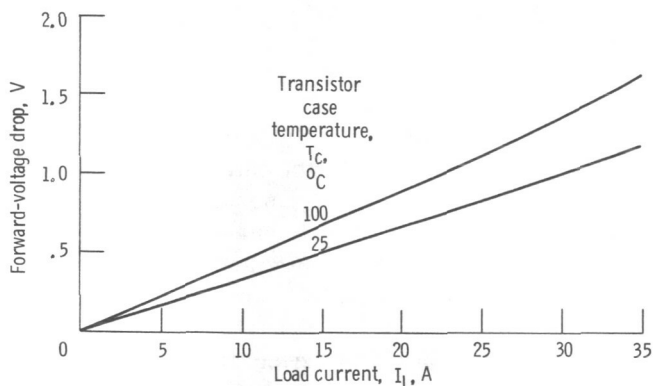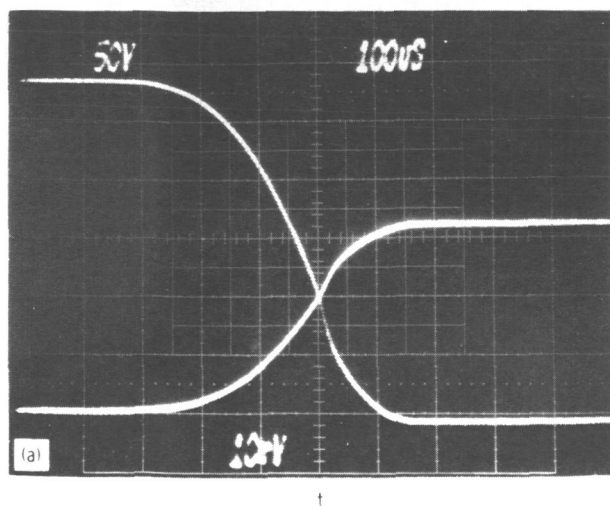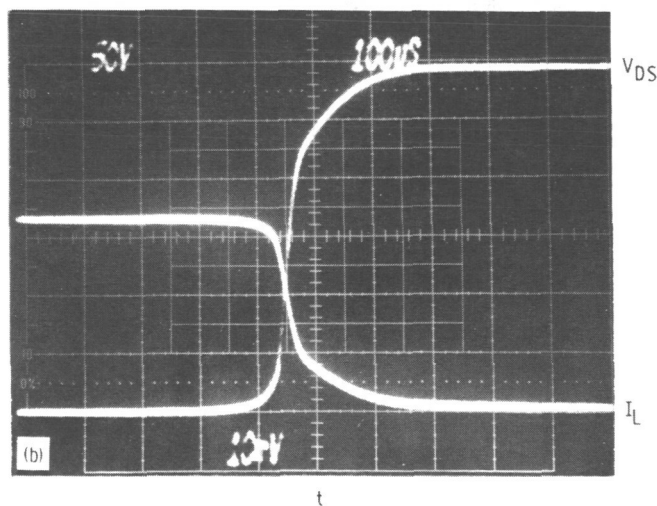
much loss, so the current must be decreased. A second consideration in picking the operating current for a given transistor is the safe operating area and the switching time. Long switching times are desirable to suppress transients and to drive capacitive loads. They also require current derating to keep operation within the safe operating area.
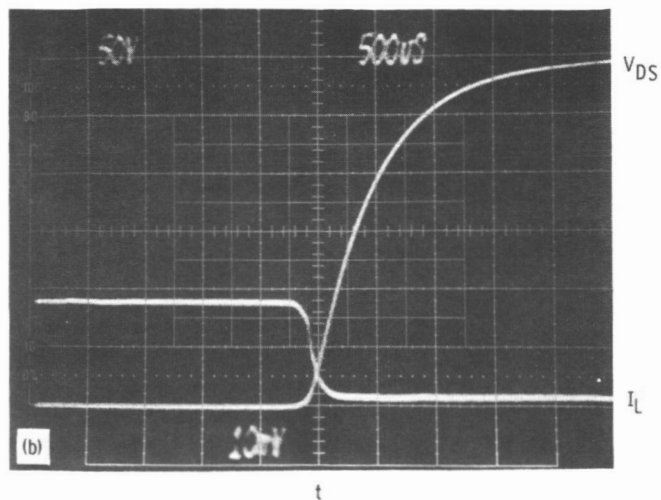
Normal turnon and turnoff.—Normal turnon and turnoff are shown in figure 10. Load conditions were 32.1 A at a voltage of 300 V. This was as close to full load current as the steps in the load bank used would allow at that voltage. Turnon time was about 300 μs and turnoff about 150 μs. These times are adequate for a wide variety of conditions.

Turnon and turnoff of a capacitive load.—Turnon of a 60-μF capacitive load at 300 V is shown in figure 11(a). The dc load for all of these tests was still 32.1 A. Note that the peak current reached 90 A, which was 10 A under the fast-trip threshold. A somewhat larger input capacitor could therefore be driven. Figure 11(b) shows turnoff under the same conditions. A capacitive load larger than one that could be successfully turned on was also tried. A 100-μF low-inductance energy storage capacitor was placed across the output. Figure 12 shows the results. The current ramped to the 100-A trip point, where the switch opened. Because of the fast tripout a short voltage transient, generated by the line inductance, appeared across the switch. It increased very rapidly to 360 V, at which point it was clipped by the protective zener diodes. It is shown that voltage transients are a limiting factor to turning on into a short circuit. The step discontinuity in the rising current waveform was caused by the differences in V_{GS} threshold among the transistors.

Turnon and turnoff of an inductive load.—Turnon and turnoff of an inductive load posed no problems so

Figure 9.—Forward-voltage drop for 35-A MOSFET remote power controller. Six VNM005A transistors in parallel.

(a) Turnon.   (b) Turnoff.

Figure 10.—Normal turnon and turnoff of a MOSFET remote power controller. Load current I_L at 10 A/division; drain-to-source voltage V_{DS} at 50 V/division; line voltage, 300 V; sweep, 100 μs/division.

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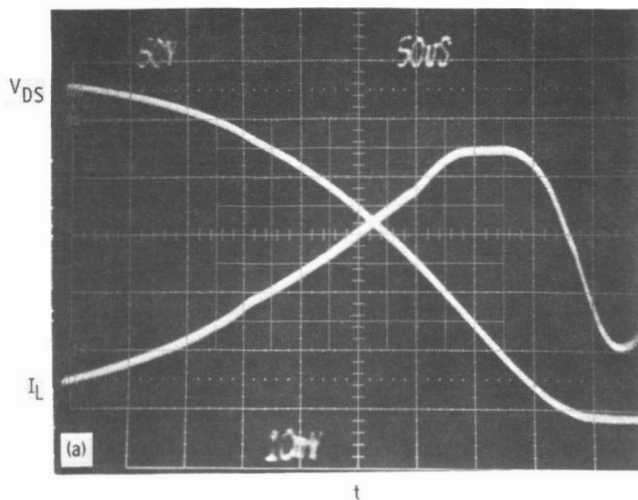
(a) Turnon. (b) Turnoff.

Figure 11.—Turnon and turnoff of a MOSFET remote power controller into 60-μF capacitive load plus full load. Load current I_L at 20 A/division; drain-to-source voltage V_{DS} at 50 V/division; line voltage, 300 V.
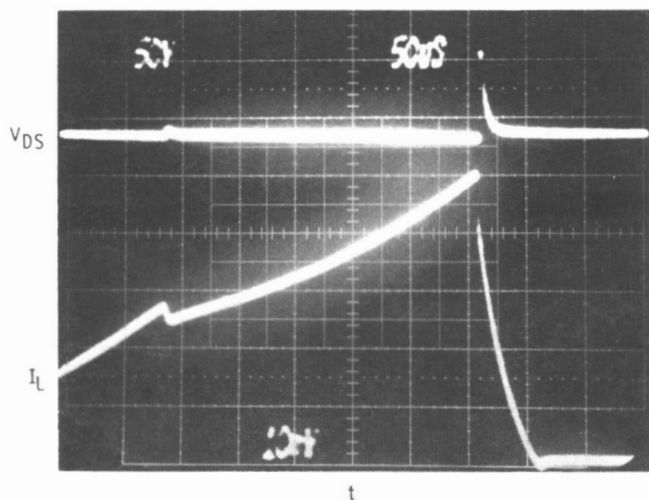

Figure 12.—Turnon of a MOSFET remote power controller into 100-μF capacitive load plus full load. Load current I_L at 20 A/division; drain-to-source voltage V_{DS} at 50 V/division; line voltage, 300 V.

long as the turnoff was normal. These conditions are shown for 100 mH in series with the previously used 32-A load in figure 13. (One would expect a smooth turnon because of the inductor regardless of the switch operation, as was the case.) The energy stored in the inductor should safely dissipate during the slow turnoff period. There was a minimal rise of 15 V in transistor voltage V_{DS} at the end of turnoff. Switching of an inductive load was very predictable and caused no undue stress on the switch.

Slow overcurrent trip.—Slow overcurrent trip is shown in figure 14. Load current is shown in a composite of trip times as the voltage was increased into an overload. Current at turnon rose to the load level, held there until the RPC tripped off, then went to zero. The apparently slow current rise at the left of the figure was not the

turnon transient of the switch but limiting caused by the limited surge current available from the test power supplies.

The lowest current that will trip the RPC is 37 A. Because tripout takes about 10 s at this level, that trip characteristic is not shown. The lowest current trip shown is about 45 A and took 2 s. As the current rose, the trip time decreased until at 100 A trip time was down to about 100 ms. At this point the fast-trip circuit took over and for higher currents tripout was instantaneous.

Fast overcurrent trip.—Fast overcurrent trip tests were run by turning on the RPC at 120 V into a load of 0.85 Ω (fig. 15). The upper trace is the current for one transistor; the lower is total load current. Individual transistor currents were very similar in shape to the total current. This particular one showed the largest deviation. Voltage across the transistors is shown in figure 16. Here the fast turnoff produced a voltage transient of just under 400 V.

Speed of turnoff is shown in figure 17. The upper trace is load current; the lower is the fast-trip comparator output V_{comp} at U3, pin 1 output. The dual-trace chopped mode was used to preserve timing accuracy. Delays in the fast-trip circuit were low enough that the current started to decrease within 1 μs of sensing the overload, and it had decreased to the one-per-unit level in 7 μs. Figure 18 shows the gate drive to the switch transistors and load current. Gate drive ramped up to the trip point, where it was suddenly clamped to ground by Q1, which was held on for the duration of the pulse from U4B. During this time U6 was recovering. At the end of the one-shot pulse period, about 400 μs, Q1 turned off. Gate drive at the output of U6 was not fully recovered to zero, increased somewhat, and then ramped to zero. This on-time of Q1 must be long enough that U6 is recovered to well below the turnon threshold of the power MOSFET's before it turns off.
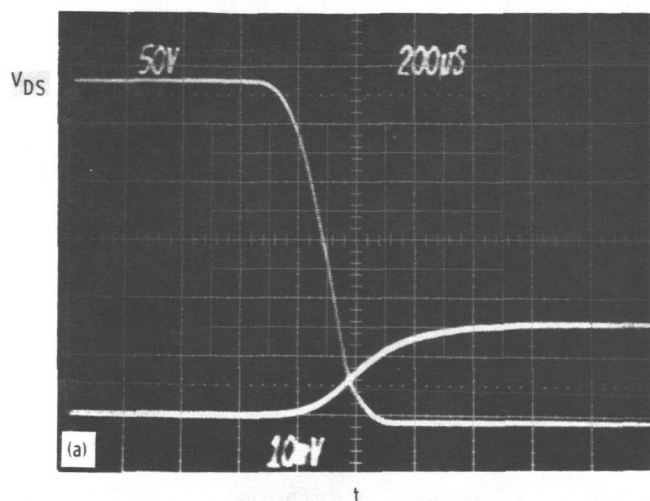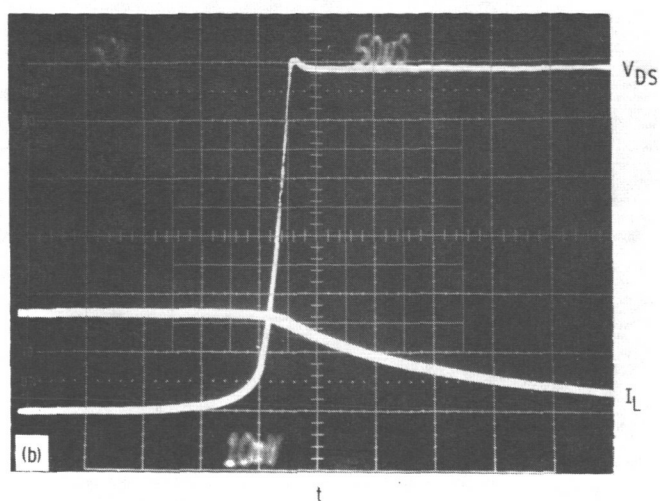
(a) Turnon. (b) Turnoff.

Figure 13.—Turnon and turnoff of a MOSFET remote power controller into a 100-mH inductive load. Load current I_L at 20 A/division; drain-to-source voltage V_{DS} at 50 V/division; line voltage, 300 V.
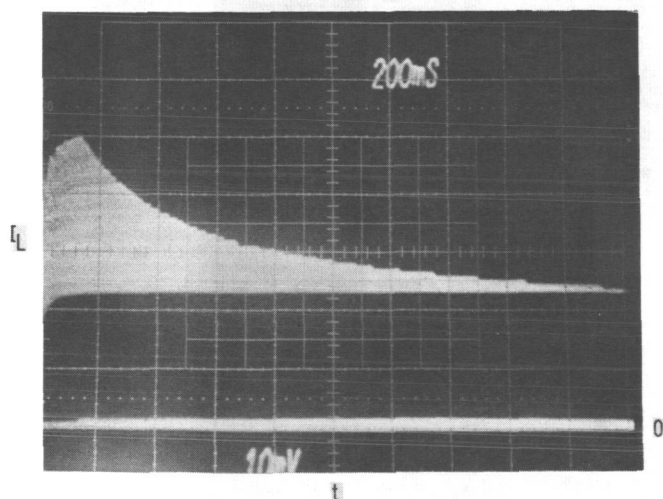

Figure 14.—Slow overcurrent trip composite characteristics of a MOSFET remote power controller. Load current I_L at 20 A/division; sweep, 200 ms/division.
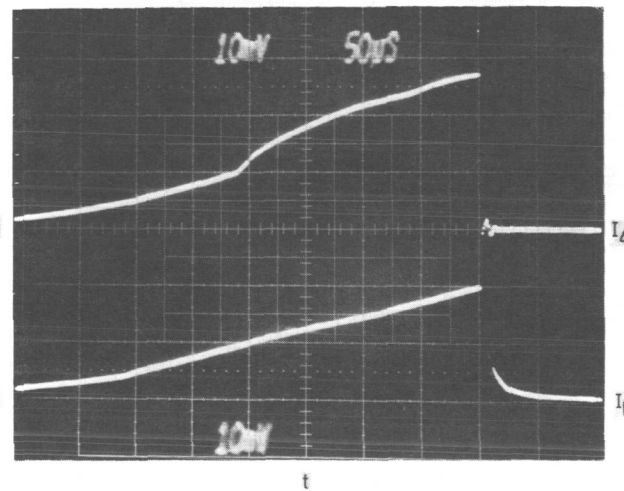

Figure 15.—Fast overcurrent trip for a MOSFET remote power controller on heavy overload during current-sharing test. Transistor 4 current I_4 at 5 A/division; load current I_L at 50 A/division; line voltage, 120 V; sweep, 50 μs/division.

Turnon into a short circuit.—Turnon into a short circuit is only slightly more stressful on the switch than other overload trips. This condition is shown for an input voltage of 300 V in figure 19. Again, note that a voltage transient was generated at turnoff that almost reached 400 V. This should not have happened because the zener surge suppressors should start to conduct at 350 V. What really happened can be seen in figure 20. This is not a short-circuit test but the switching of 300 V into 1.85 Ω, which is a severe overload. The transistor drain-to-source voltage V_{DS} shown in the lower trace increased rapidly, peaked near 400 V, and about 1/2 μs later was clipped back to the 350-V level by the protective zeners. One reason for the first peak was lead inductance between the switching transistors and the protective zeners.

Relocating the zeners closer to the transistors would eliminate or at least decrease this portion of the transient.

To check what would happen in a more normal application, the RC network at the RPC input was disconnected and a short-circuit test was performed (fig. 21). With a line voltage of only 100 V the turnoff transient produced a voltage transient that reached 380 V. In about 1/2 μs this was clipped to the 350-V zener level. Using the values from this figure of 280 V for the transient and 40 A/μs for di/dt indicated that the circuit inductance was 7 μH. This is a reasonable value for power systems on a satellite where there are no long cable runs.

Short circuit from full load.—Short circuiting the output from a full-load condition was the final test run
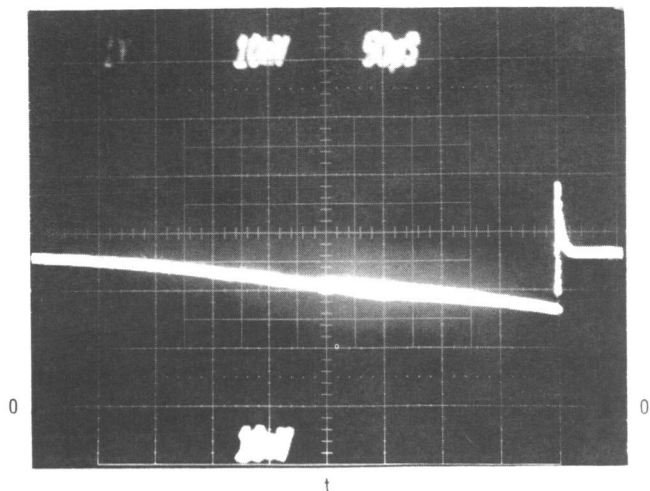
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Figure 16.—Fast overcurrent trip voltage spike for a MOSFET remote power controller. Drain-to-source voltage V_{DS} at 100 V/division; line voltage, 275 V; load resistance, R_L, 0.85 Ω; sweep, 50 μs/division.
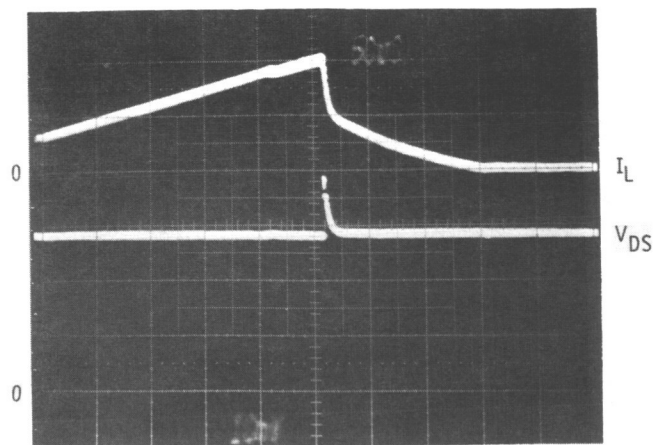

Figure 19.—Turnon of a MOSFET remote power controller into a short circuit. Load current I_L at 50 A/division; drain-to-source voltage V_{DS} at 100 V/division; line voltage, 300 V; sweep, 50 μs/division.
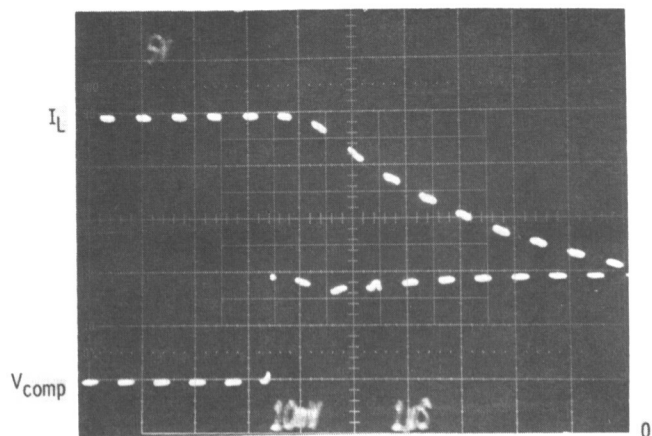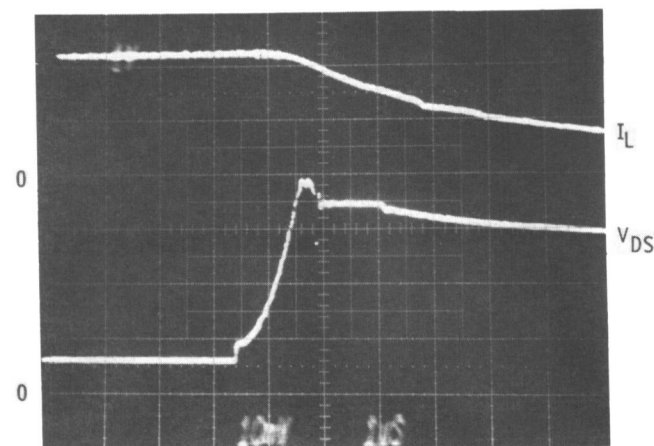

Figure 17.—Fast overcurrent trip turnoff delay for a MOSFET remote power controller. Load current I_L at 20 A/division; comparator output V_{comp} at 5 V/division; line voltage, 120 V; load resistance, R_L, 0.85 Ω; sweep, 1 μs/division.

Figure 20.—Turnon of a MOSFET remote power controller into a heavy overload. Load current I_L at 50 A/division; drain-to-source voltage V_{DS} at 100 V/division; line voltage, 300 V; load resistance, R_L, 1.85 Ω; sweep, 1 μs/division.
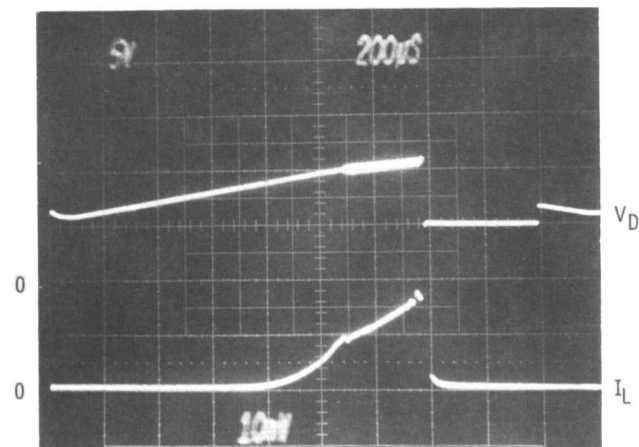

Figure 18.—Fast overcurrent trip gate drive for a MOSFET remote power controller. Load current I_L at 50 A/division; drain voltage V_D at 5 V/division; line voltage, 275 V; load resistance, R_L, 0.85 Ω; sweep, 200 μs/division.
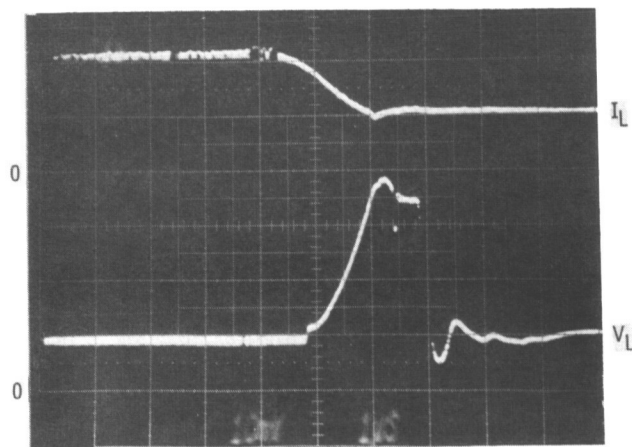

Figure 21.—Turnon of a MOSFET remote power controller into a short circuit with no added capacitor at RPC input. Upper curve: load current I_L at 50 A/division. Lower curve: load voltage V_L at 100 V/division. Line voltage, 100 V; sweep, 1 μs/division.

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on the RPC. For this test a large SCR was connected across the load bank. After the load was adjusted to 35 A at 270 V, the SCR was fired. The waveforms for V_{DS} and I_L (fig. 22) are similar to those for other overload cases.

Thermal vacuum tests.—Thermal vacuum testing was performed in a 40.6-cm (16-in) bell jar vacuum system with a turbomolecular pump. The RPC was mounted on a temperature-controlled baseplate and operated through the basic turnon and turnoff tests at various voltage and current levels. No problems were encountered and the RPC functioned properly. Vacuum level in the bell jar was about 10^{-5} torr, sufficiently low that no heat could be transferred by convection. Temperatures were measured at the heat sink to which the GTO is directly mounted, the RPC base, and the temperature-controlled baseplate to which the RPC was bolted. Runs at three temperatures were made to demonstrate the adequacy of the thermal design. Data from these tests are shown in table IV. Of the three tests the largest temperature difference between heat sink and mounting plate was 7.3 deg C at a mounting plate temperature of 49.6 °C.

Gate Turnoff Thyristor (GTO) RPC's

Two limitations of MOSFET switches are the maximum breakdown voltage rating and the current that can be switched by a single transistor. GTO devices operate at higher current densities and are available with much higher single-device current ratings. They also have voltage ratings to 2500 V. GTO devices were first introduced at low power levels in the 1960's. They did not find wide application until recently. In 1978 higher power devices were developed by G.E., and several foreign manufacturers followed soon thereafter. Only in the last year or two have American manufacturers taken this market seriously. There is currently a good variety of

Figure 22.—Short circuit of a MOSFET remote power controller from full load. Load current I_L at 20 A/division; transistor voltage V_{DS} at 50 V/division; line voltage, 270 V; sweep, 2 μs/division.

TABLE IV.—MOSFET RPC (MODEL E) THERMAL VACUUM TEST DATA ON IRF350 TRANSISTORS FOR THREE MOUNTING PLATE TEMPERATURES

	Test		
	1	2	3
Mounting plate temperature, °C	79.9	49.6	25.7
Load current, A	34.87	34.82	34.86
Load voltage, V	172.7	172.3	172.4
MOSFET on-voltage, V	2.61	2.14	1.80
MOSFET heat sink temperature, °C	87.0	56.9	31.8
Baseplate temperature, °C	82.6	52.2	27.9

GTO's available for switchgear use. A representative number are listed in table V.

Some of the newer devices are even available in cases with internal isolation. Examples are the Westinghouse GDM series and the Telefunken G50A. None were available when the RPC's to be described were being designed. They would simplify construction for any new designs and minimize thermal resistance. From the table it can be seen that large enough GTO's are available that the need to parallel them should not occur. The only limitation on handling large currents will be the physical size of the package.

Advantages and Disadvantages of GTO Devices

RPC's to switch dc have been developed using SCR's (refs. 9 and 11). Although having advantages such as high voltage and current capability, they are difficult to turn off. A commutation circuit that can reduce the SCR current to zero until it recovers blocking capability is required. The circuit must store and process considerable energy by using bulky capacitors, inductors, and transformers. Additionally, a recharge circuit to provide the energy at the proper voltage must be provided. The net result is that the turnoff circuitry conservatively doubles the weight and volume of an SCR RPC.

Gate turnoff thyristors have all of the advantages of SCR's and fewer disadvantages. Specifically, the advantages include

(1) Simple, low-power turnon

(2) No holding power required

(3) High voltage and current ratings

(4) Turnoff by low-energy pulse (high current, low voltage, and short duration)

(5) Forward-voltage drop with small negative temperature coefficient

(6) Microsecond turnoff times

For use in RPC's, the GTO also has a few disadvantages:

(1) Forward-voltage drop of 2 to 3.5 V

(2) Minimum holding current required for it to stay on

(3) Turnon and turnoff times not controllable

Forward-voltage drop will probably be a limitation only for low- to medium-voltage switches. Above 600 V the
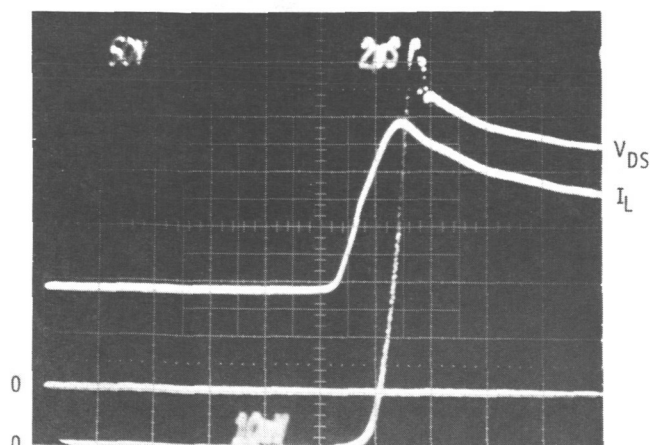
TABLE V.—REPRESENTATIVE GATE
TURNOFF THYRISTORS

Device	Repetitive peak off-state voltage, V	Average on-state current, A	Controllable anode current, A
Amperex:			
BTV59-1000R	1000	15	50
BTW58-1500R	1500	5	25
AGE Telefunken:			
G20A	1200	6.5	20
G50A	1200	14	50
G90A	1200	27	90
Hitachi:			
GFT50A6	600	20	50
GFF90A6	600	42	90
GFF200E12	1200	70	200
International Rectifier (IR-160PFT120)	1200	160	600
Sony:			
SG6574A	1000	15	75
SG6563	1000	40	160
Toshiba:			
SG800 R21	1300	400	800
SG1400 EX21	2500	700	1400
Westinghouse:			
GDM11209	1200	20	90
GDM11220	1200	45	200
GSD11245	1200	125	450

percentage loss decreases to acceptable levels. Another mitigating factor is the negative temperature coefficient of the forward-voltage drop, particularly when compared with MOSFET's. It is not necessary to derate the current rating of a GTO that will be operated at high temperatures to compensate for forward drop, as must be done for MOSFET's.

The short transition times of the GTO can be both an advantage and a disadvantage. A fast and uncontrollable turnon time precludes switching of capacitive loads unless there is some current limiting in the power system or the capacitance is small. For normal turnon and turnoff, slower transition times would be desirable, as was explained in the MOSFET section. With the GTO this is not possible. The turnon and turnoff times are characteristic of the device and cannot be appreciably modified by the drive circuitry. This is the one major drawback of GTO's and does somewhat limit the range of applications for RPC's using them. On the other hand, the fast turnoff is a necessity for successfully disconnecting heavy overloads or short circuits before catastrophic damage occurs. GTO's do not turn off as

fast as MOSFET's, but the 2- to 10-μs turnoff time is fast enough for most applications. Transient voltage generation at turnoff is similar to the problem with MOSFET's. Since GTO's are available with high voltage ratings, the magnitude of this voltage is less of a problem. Because high dv/dt at turnoff may cause the GTO to retrigger on, this becomes a design consideration.

GTO RPC Design Philosophy

Considering the limitations of GTO's, it is clear that RPC's using them will have somewhat different applications than MOSFET RPC's. GTO RPC's will be suitable for high-voltage, high-power applications, particularly where the power source is somewhat current limited. A good example would be high-voltage solar array switching. If the source has little or no current limiting, such as a power system using batteries, GTO RPC's may still be practical if wiring inductance is sufficient to limit initial current rise time. The RPC must be matched to the application.

The forward-voltage drop of a GTO is very similar to that of an SCR, but a bit higher. It is also higher than the drop normally attained with MOSFET's. To have comparable efficiency, GTO switches would therefore have to operate at higher voltages. This disadvantage is partially offset by the slightly negative temperature coefficient of forward voltage. Instead of a large increase at high temperature, as with MOSFET's, the forward-voltage drop decreases slightly, partially compensating for the higher starting value. As might be expected, the forward-voltage drop is also lower for GTO's with lower anode voltage ratings.

Turnon requirements for a GTO are very similar to those for an SCR. Increasing gate current decreases turnon delay but does not appreciably change the rise time. It is important to provide sufficient current to fully turn on the GTO. If this is not done, some regions of the GTO may not turn on, and this will cause destructive currents in those regions that did turn on. It is also necessary to maintain gate current until the anode current has increased beyond the latching current. If this is not done, the GTO will attempt to turn off, and this can result in destructively high current densities in the remaining on-regions.

The major advantage of the GTO over the SCR is that it can be turned off by applying a negative current pulse to its gate terminal. This eliminates the need for a large, high-powered, forced commutation circuit. The negative gate current required to turn off a GTO is typically about one-fifth of the anode current at a voltage of 8 to 15 V. It need only be applied until the GTO is fully turned off, typically less than 10 μs. Because of the short turnoff time the power required is small and the energy can be stored in a capacitor. This eliminates the need for a high-

current, isolated power supply, which would take more space and add weight to the design. It would also have more power loss, which would degrade efficiency.

Turning off a GTO switch from a normal on-condition requires only that sufficient negative gate current be supplied. Several turnoff circuits are described in the manufacturers' literature. All are suitable for normal turnoff because there are no severe time constraints. If there is some inductance in the system, a transient voltage proportional to di/dt will be produced. It must be limited by using zener suppressors. The transient voltage produced may also be a problem. If it rises too fast, it can exceed the dv/dt rating of the GTO and cause it to retrigger on. A snubber circuit must be used to prevent this.

Turning off a GTO from an overload or short-circuit condition is far more demanding and is the most significant design consideration. For an RPC to function properly under all types of overloads, including short circuits, it must be self-protecting. This requires sensing the overcurrent and turning off before the current can rise to a level where control is lost. The worst case is a system where the power controller is connected to a low-impedance bus with considerably more current capability than the rating of the controller. Short-circuit current will rise at a rate controlled by the system inductance. In this case circuit inductance is very desirable. It would be advantageous to add inductance, but for most cases an inductor large enough to slow the current rise sufficiently would be as large and heavy as the whole RPC. The most reasonable alternative is to design the overcurrent sense-and-turnoff circuit to have the fastest possible response. This is not easily accomplished as a number of tradeoffs are involved.

Most turnoff circuits recommended by GTO manufacturers incorporate a small inductor in series with the gate. Its purpose is to maintain current flow out of the gate as its impedance increases during turnoff. This inductor also limits the initial buildup of negative gate current and slightly delays the onset of turnoff. If the gate inductor is not used, sufficient negative gate drive may not be maintained long enough and the GTO may retrigger on. This can be due to feedback coupling to the gate from the very high dv/dt generated by the fast turnoff. A compromise approach taken with the GTO RPC's built was to drive the gate off from a low-impedance source by using power MOSFET's to switch extremely rapidly. A minimum inductance, less than 1 μH, was also used.

Capacitive storage provided the turnoff energy. Tantalum electrolytic capacitors were originally used. They were capable of storing the required energy in minimal volume but limited the rise time of the turnoff pulse. Replacing them with plastic or ceramic capacitors of lower value actually improved turnoff by decreasing the rise time. Several capacitors were paralleled to minimize lead inductance and resistance. To allow more margin for turnoff, the trip threshold was set low, at two per unit. This gave the turnoff circuit a little more time to act before the anode current could exceed the device turnoff capability.

Turning off the GTO fast enough to prevent failure when it is short circuited would be difficult enough if this were the only consideration. In practice an additional problem occurs. When the GTO turns on, the very fast normal current rise time couples an undesirable signal into the trip circuit. This signal tends to trip off the RPC and may do so. When this happens, the RPC appears to be ignoring the turnon command. What is really happening is that the GTO turns on and in the process couples enough of a transient into the fast-trip circuit to turn off. The whole process takes less than 5 μs. The only solution to this problem, once the circuit is built, is to slow the response of the fast-trip comparator by adding a capacitor across its input. This successfully prevents the spurious trip point during turnon. It also slows the response for a short-circuit trip to the extent that the RPC may not respond fast enough under worst-case conditions. A better cure would be to change the circuit layout to minimize the undesirable feedback coupling.

Controls and Fabrication

Controls for GTO RPC's can be very similar to those for MOSFET RPC's. Specifically the same ground-referenced control and power supply card is used. The only difference is in the level and number of isolated power supplies provided. The floating control portion of the circuit is also very similar. Overload sensing circuits are identical. Turnon and turnoff circuits are similar but produce narrow drive pulses instead of latching on or off. A third booster card is also required for the higher powered GTO's to provide the high turnoff current. The circuits built are quite flexible and can be adapted to almost any GTO by scaling the drive pulse magnitudes.

As was done with the MOSFET RPC's, these units (fig. 23) were packaged in a flight-qualifiable configuration but not fully tested for shock and vibrations. All units have drawn-aluminum covers, which were removed to show construction details.

Specifications of GTO RPC's Built

GTO switches were used in four of the RPC's. Unit B, which was developed to switch positive and negative power to an ion engine, used GTO's in both sections. This was the most practical way of providing the necessary high-voltage capability. Units C and D are general-purpose RPC's, and their design is generally applicable to GTO RPC's of any rating with appropriate modification for gate drive required. Specifications for the four RPC's are given in the table VI.
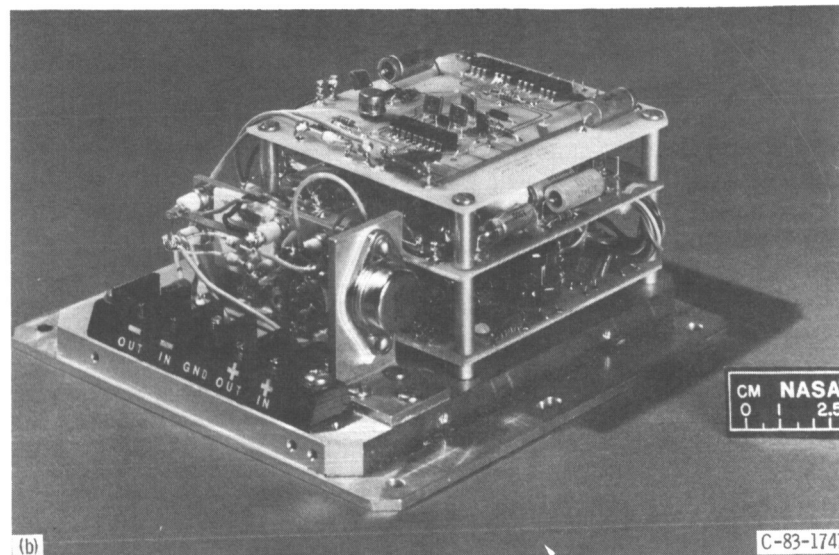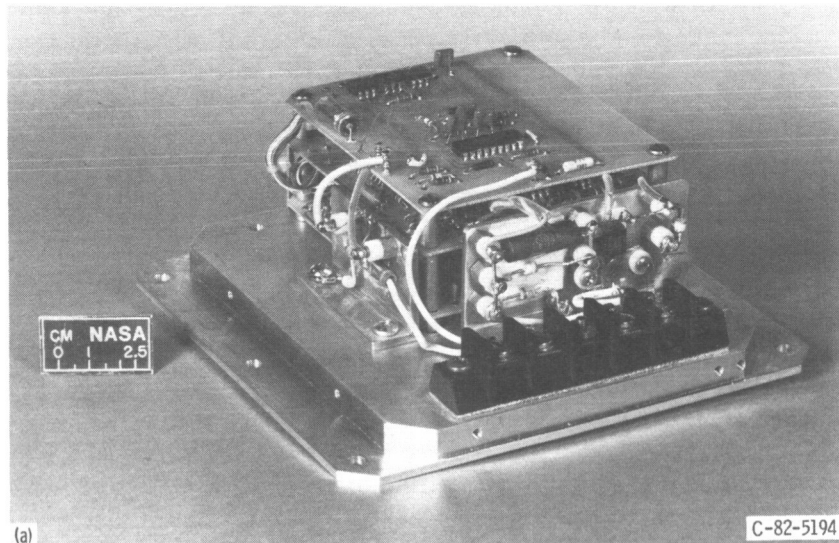
(a) Type A.   (b) Type B.

Figure 23.—Gate-turnoff-thyristor remote power controllers.

GTO RPC Circuit Description

All of the RPC's use the ground-referenced control circuit previously described in the MOSFET RPC section. For the two low-power units, A and B, the complete floating control circuit for one switch is contained on one printed-circuit card. For the higher powered units, C and D, the large turnoff current required necessitated an additional power driver card, on which is mounted the energy storage capacitors. The latter is the top card visible in figures 23(c) and (d). The 35-A type C RPC described herein is very similar in circuitry to all the other GTO units and has the same current rating as the MOSFET unit described previously.

Detailed circuit descriptions are given by circuit function and referred to the detailed electrical schematic (fig. 24).

Normal turnon from the ground-referenced control board is by optoisolator OP1. Its output clocks a one-shot multivibrator, U7A, that generates a 10-μs, negative-going pulse. This pulse directly turns on MOSFET Q1, which connects capacitor C14, charged to +10 V, to the gate of the GTO through a 10-Ω resistor and turns it on. Because of some resistance in transistor Q1, the actual turnon gate current is 0.7 A.

Normal turnoff is similarly coupled via optoisolator OP2. It clocks multivibrator U7B to produce a 15-μs negative-going pulse at pin 8, which turns on comparator U5. Output of the comparator is normally at −10 V.
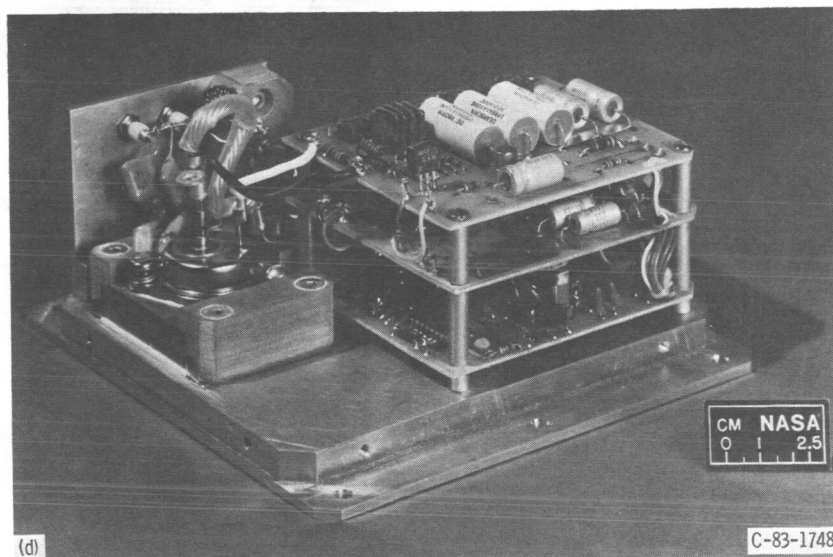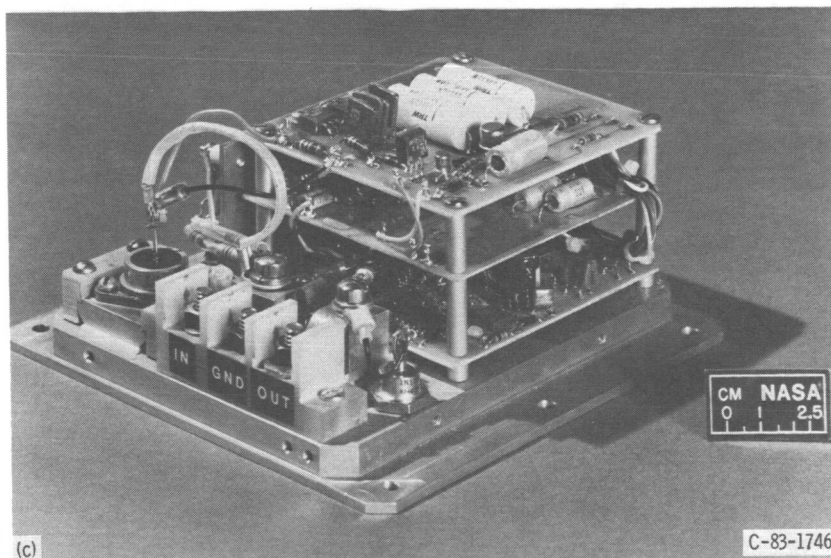
(c) Type C (35 A at 800 V dc).   (d) Type D (65 A at 800 V dc).

Figure 23.—Concluded.

TABLE VI.—SPECIFICATIONS OF GTO RPC's

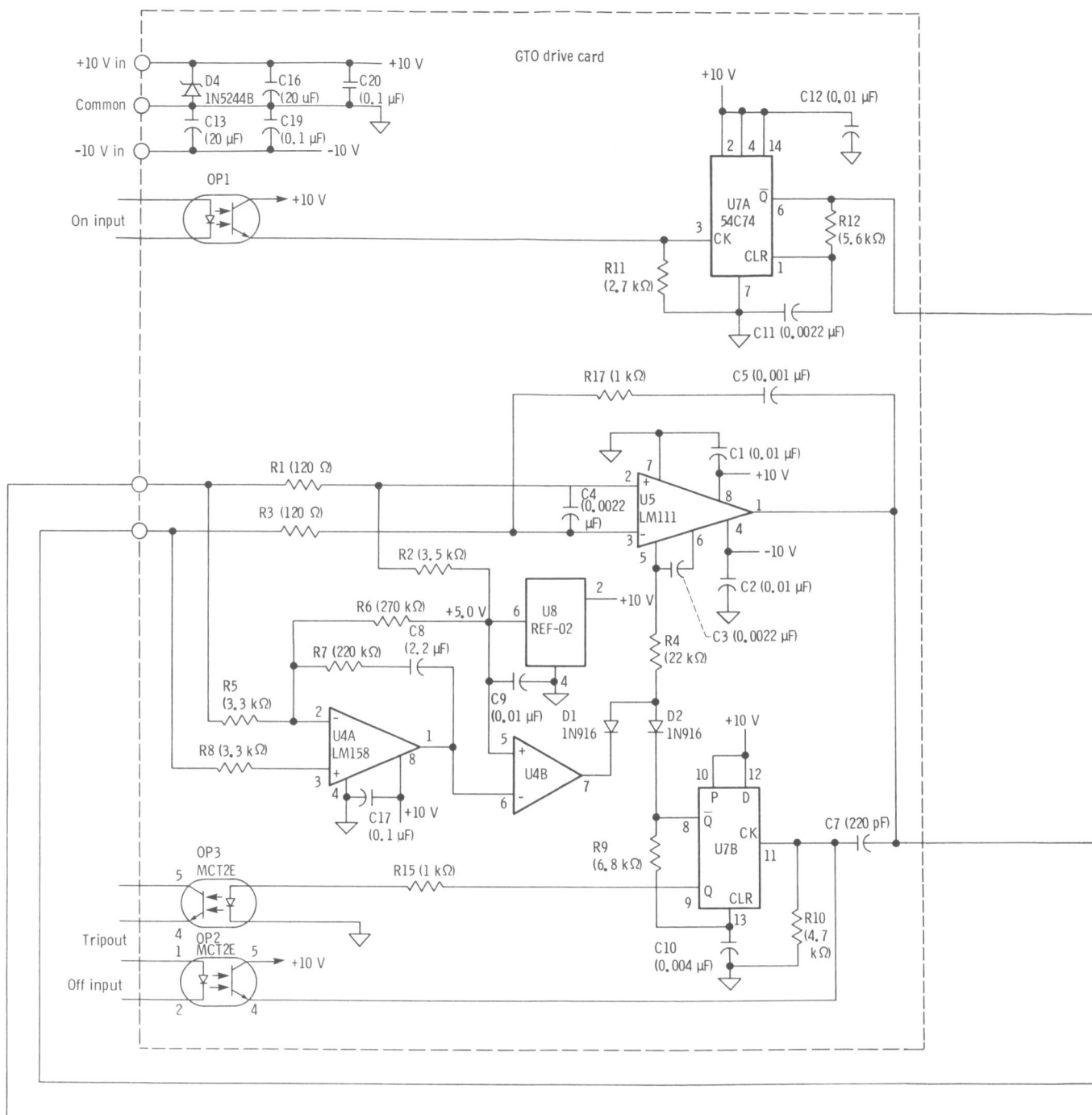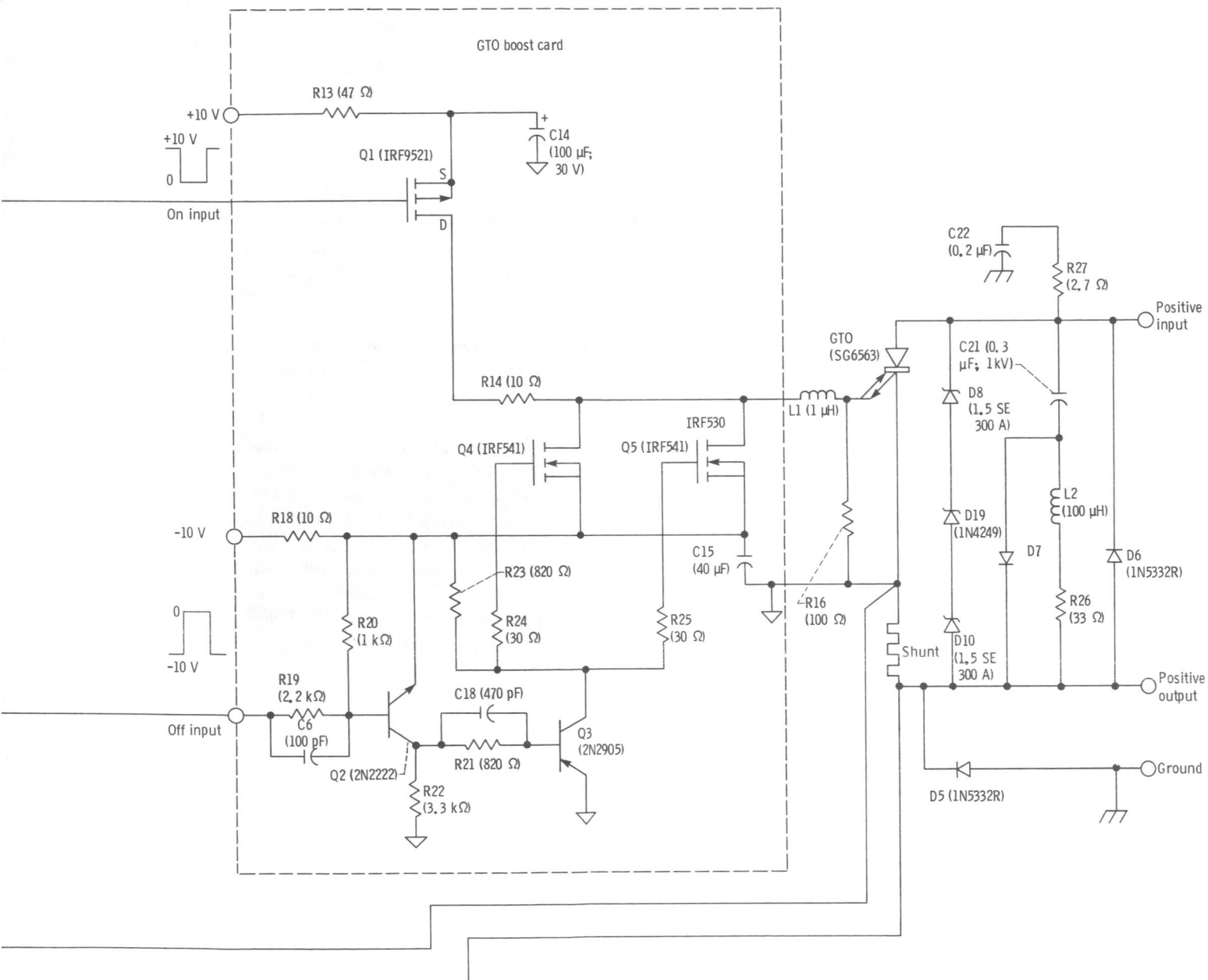|	RPC model				
	A	B		C	D
		Positive section	Negative section		
Operating voltage, V	1200	1000	−1100	800	800
Full load current, A	6.5	15	0.1	35	65
Switching time (on/off), μs	2	8	2	10	8
Switch-on drop, V	3.3	2.1	1.8	2.4	2.1
Efficiency, percent	99.7	99.7	99.8	99.7	99.7
Device	BTW59	SG6574A	BT157	SG6563	GFF200E12
Weight, kg	1.4	1.5		1.7	2.2

Figure 24.—Schematic diagram of

GTO boost card

- +10 V — R13 (47 Ω)
- +10 V / 0 — On input
- Q1 (IRF9521) — S, D
- C14 (100 μF; 30 V)
- R14 (10 Ω)
- Q4 (IRF541)
- Q5 (IRF541) — IRF530
- L1 (1 μH)
- GTO (SG6563)
- C22 (0.2 μF)
- R27 (2.7 Ω)
- Positive input
- C21 (0.3 μF; 1kV)
- D8 (1.5 SE 300 A)
- -10 V — R18 (10 Ω)
- R23 (820 Ω)
- R24 (30 Ω)
- C15 (40 μF)
- R25 (30 Ω)
- R16 (100 Ω)
- D19 (1N4249)
- D7
- L2 (100 μH)
- D6 (1N5332R)
- 0 / -10 V
- R20 (1 kΩ)
- R19 (2.2 kΩ)
- C6 (100 pF)
- Off input
- C18 (470 pF)
- Q3 (2N2905)
- Q2 (2N2222)
- R21 (820 Ω)
- R22 (3.3 kΩ)
- Shunt
- D10 (1.5 SE 300 A)
- R26 (33 Ω)
- Positive output
- D5 (1N5332R)
- Ground

35-A GTO remote power controller.

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When it is turned on, its output is a positive-going pulse rising to zero volts. This pulse is buffered by transistors Q2 and Q3 to provide a low-impedance drive for MOSFET's Q4 and Q5. When turned on, they connect capacitor C15, which is charged to -10 V, to the gate of the GTO. Capacitor C15 is made up of several Mylar or ceramic capacitors in parallel and can supply a turnoff pulse of more than 30 A with a rise time of 0.5 μs. Should a turnoff signal occur before the turnon pulse has disappeared, as would happen if the RPC were turned on into a heavy overload, the turnoff circuit overpowers the on-drive. Resistor R14 limits the current in this situation.

Overcurrent tripping uses the same basic sensing circuit as the MOSFET RPC. Operational amplifier U4A "integrates" the current signal. U4B is used as a comparator and turns on the U5 comparator via D1 and R4 when the integrated trip signal exceeds 5 V. From this point on, the circuit operates the same as for normal turnoff.

The fast-trip circuit, which uses comparator U5, senses when the current has exceeded twice normal and immediately shuts down the RPC. The trip threshold is set by R2, which provides a bias from precision 5-V reference U8. When this positive bias is exceeded by the negative signal from the current shunt, the comparator is triggered. To ensure that it remains in the tripped state long enough for the GTO to fully turn off, two feedback circuits are used. One consists of R17 and C5 and provides positive feedback around the comparator to hold it on until the second loop can operate. When U5's output goes positive, it triggers U7B via capacitor C7. This generates a 15-μs pulse, the same as a commanded turnoff, which holds the comparator on via D2 and R4. In this way the turnoff pulse is always maintained for 15 μs.

The switch portion consists of the GTO and its protective circuitry and snubber network. This is a critical part of the circuit for fast-trip operation. In normal operation, snubber capacitor C21 is discharged through L2, R26, and the GTO when the latter turns on. Discharge time is determined by L2 and R26. It is important that C21 is discharged quickly so it can absorb energy should the GTO turn off soon after turnon. Diodes D5 and D6 protect against reverse-voltage transients, and diodes D8, D9, and D10 are a zener clamp to limit positive transients.

Experimental Evaluation of 35-A GTO RPC

Each of the GTO RPC's was tested over the full range of operating conditions. They were tested both on the bench and under thermal vacuum conditions. The various tests performed included normal turnon and turnoff, static forward-voltage drop, turnon and turnoff of an inductive load, turnon and turnoff of a capacitive load, slow overcurrent trip, fast overcurrent trip, turnon into a short circuit, short circuit from full load, and thermal vacuum testing. Test conditions were almost identical to those used in evaluating the MOSFET switches. This was done on purpose to make possible meaningful comparisons between the two switch types, which have identical current ratings.

Normal turnon and turnoff.—Normal turnon and turnoff at full rated current and voltage are shown in figure 25. Test conditions were identical to those used with the MOSFET RPC except that the filter capacitor added at its terminals was not used. Without this capacitor the switch sees the full line and load inductance. It was calculated to be 7 μH from current rise-time data. Note in figure 25 that this inductance caused a 160-V overshoot and ringing at turnoff.

Static forward-voltage drop.—Static forward-voltage drop and latching current were also measured. The test voltage was varied to change operating current in exact 5-A steps (fig. 26). Latching current, the minimum current that the GTO must carry to remain on, was 2.8 A under the same test conditions.

Turnon and turnoff of an inductive load.—Turnon and turnoff of a 100-mH inductive load are shown in figure 27. Current rise and fall times were much longer than for a resistive load, and the turnoff transient was not much worse except that it lasted longer.

Turnon and turnoff of a capacitive load.—Turnon of a capacitive load is a more difficult condition for the RPC. It was tried with only 0.5 μF in parallel with 24 Ω at a supply voltage of 300 V. The RPC would not turn on. Figure 28 shows the current and voltage at turnon. The rate of current rise was very rapid, but the peak current reached was only 35 A. It was well below the set trip point, an indication that the fast-rise-time pulse was being unintentionally coupled back to the trip circuit.

Slow overcurrent trip.—To display the slow overcurrent trip characteristic of the RPC, it was set up at low voltage with a resistive load current slightly greater than the continuous rated current. Increasing the voltage after each attempt at turning on then decreased the trip time. A composite photograph (fig. 29) shows the current. Threshold was just above 35 A, at which current tripout would take several seconds. At 40 A, the lowest trace in the figure, tripout would still take more than 2 s. As the current rose, trip time smoothly decreased until at 65 A it only took 50 ms to trip. Any larger current caused a fast, or almost instantaneous, tripout.

Fast overcurrent trip.—Fast overcurrent trip tests were run at 375 V with a load resistance of 4.8 Ω. Current and voltage waveforms are given in figure 30(a). Peak current was about 65 A. Some unintentional feedback due to the very fast current rise reached the comparator and lowered the trip threshold slightly. Also a fast voltage spike generated at turnoff exceeded 800 V for the moderate line voltage used. Inability to control the dv/dt of this transient voltage was a severe limiting factor in RPC
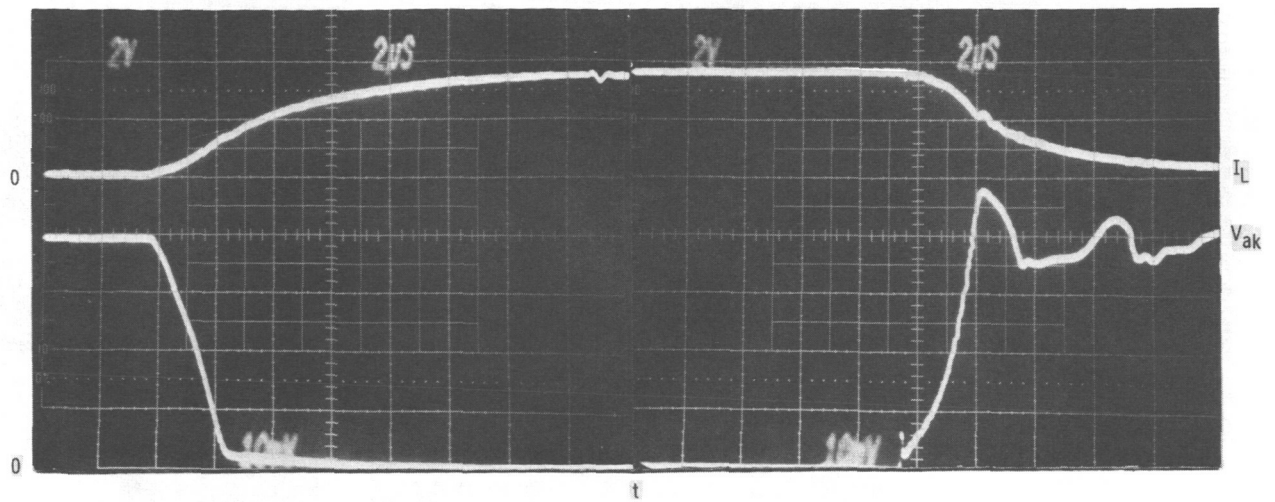
Figure 25.—Normal turnon and turnoff of a type C GTO remote power controller. Load current I_L at 20 A/division; voltage across GTO V_{ak} at 200 V/division; line voltage, 800 V dc; load, 35 A.
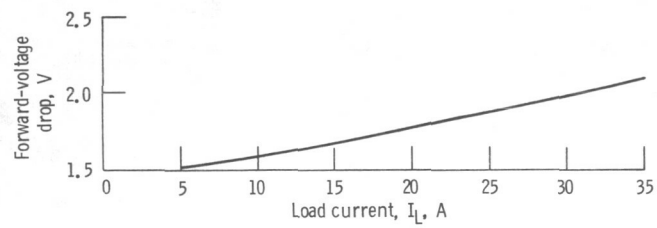

Figure 26.—Forward-voltage drop for 35-A GTO remote power controller. Sony SG6563. Case temperature, 25 °C.
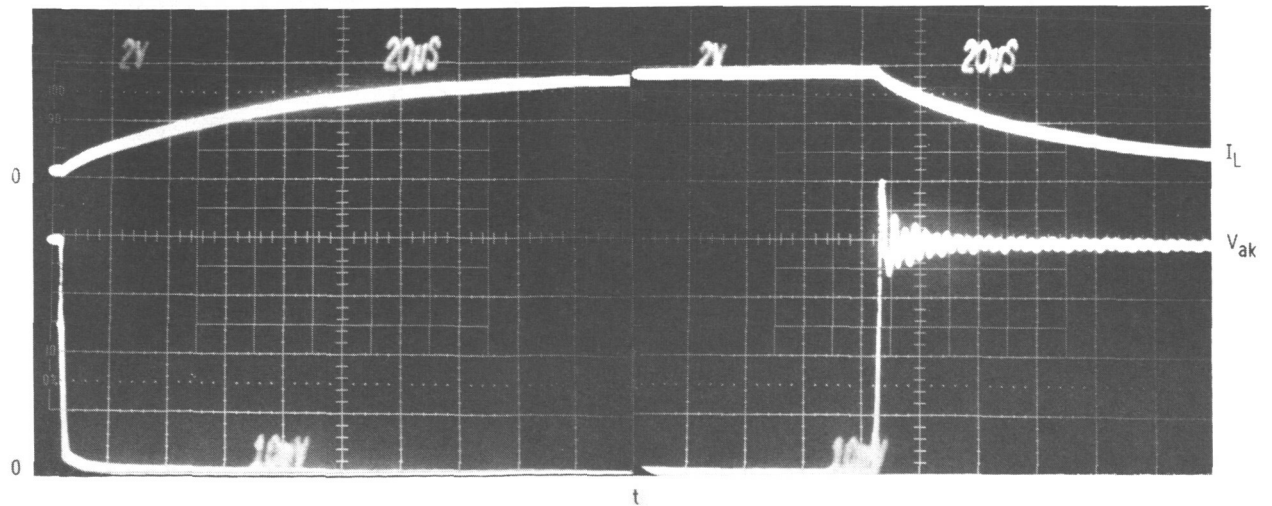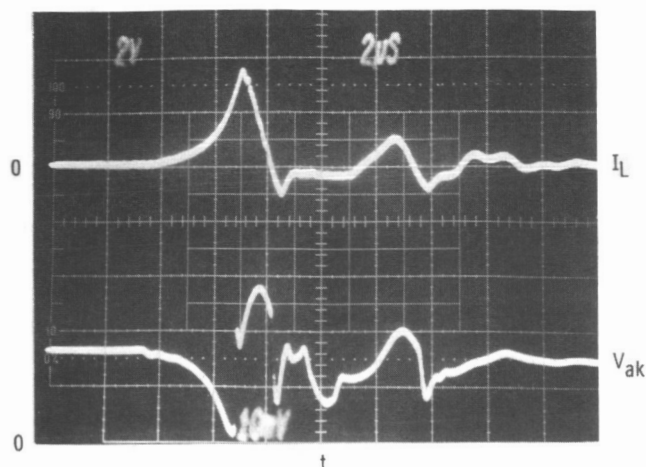

Figure 27.—Turnon and turnoff of a GTO remote power controller into a 100-mH inductive load. Load current I_L at 20 A/division; voltage across GTO V_{ak} at 200 V/division; line voltage, 800 V dc.

Figure 28.—Turnon of a GTO remote power controller into a 5-μF capacitive load. Load current I_L at 20 A/division; voltage across GTO V_{ak} at 200 V/division; line voltage, 300 V; load resistance, R_L, 24 Ω.
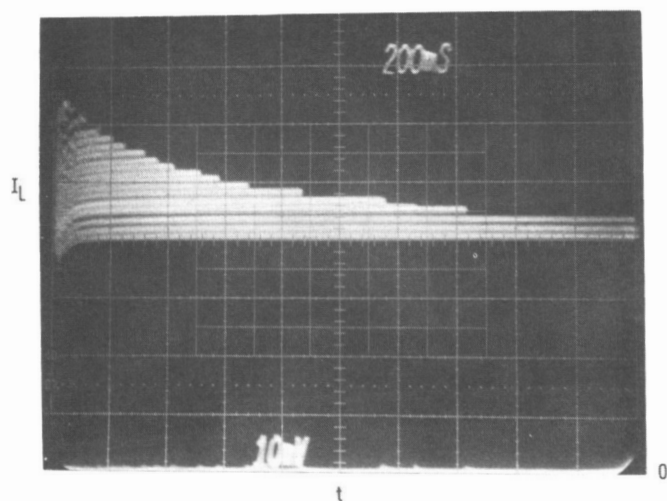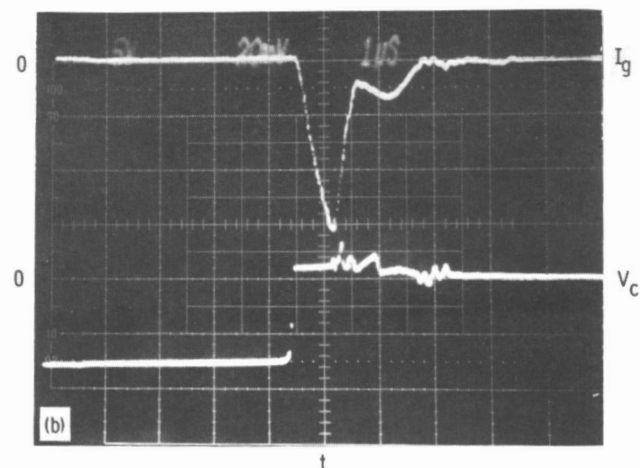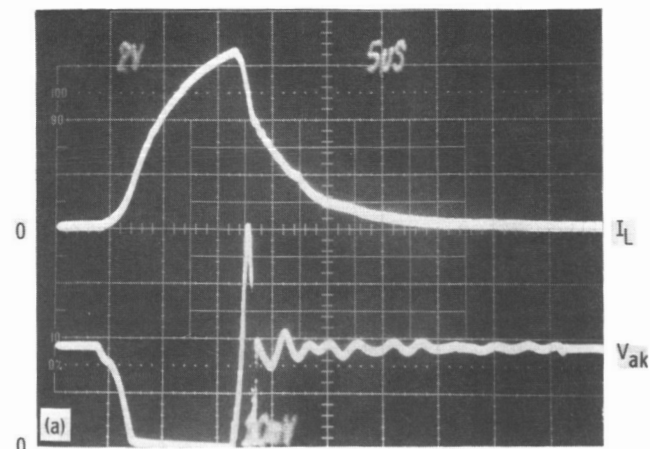

Figure 29.—Slow overcurrent trip composite characteristic of a GTO remote power controller. Load current I_L at 10 A/division.

(a) Load current I_L at 200 A/division; voltage across GTO V_{ak} at 200 V/division; sweep, 5μs/division.
(b) Gate current I_G at 10 A/division; LM111 output voltage at 5 V/division; sweep, 1μs/division.

Figure 30.—Fast overcurrent trip for a GTO remote power controller.

design. Some limiting was possible with an RC network added from output to ground. It slowed the dv/dt and prevented it from turning the GTO back on. Figure 30(b) shows the speed of response of the turnoff circuit. When the fast-trip comparator (U5) sensed the overcurrent, its output went high. Within 0.1 μs the turnoff gate drive started; it reached peak off-current in less than 1 μs.

Turnon into short circuit.—Turnon into a short circuit was modified somewhat in that a full short was not used. Instead a load resistance of 0.85 Ω, which simulates most real life shorts adequately, was used. Current and voltage for this condition are shown in figure 31.

Thermal vacuum tests.—The GTO RPC's were thermal vacuum tested in the bell jar facility described previously. Data from the type C test at three baseplate temperatures are shown in table VII. For these conditions the largest

temperature difference between heat sink and mounting plate was 25.6 deg C at a mounting plate temperature of 80.4 °C. Of the various GTO designs, this one had the highest thermal drop.

Short circuit from full load.—Short circuit from full load (fig. 32) was performed by turning on a large SCR connected across the load bank. Since the GTO was already fully conducting, the current rose rapidly at about 20 A/μs. The RPC was operating at full load and the fast-trip comparator was at about half threshold. When the current rose, the comparator tripped quickly and turned off the GTO in 2 μs.

Hybrid Packaging Option

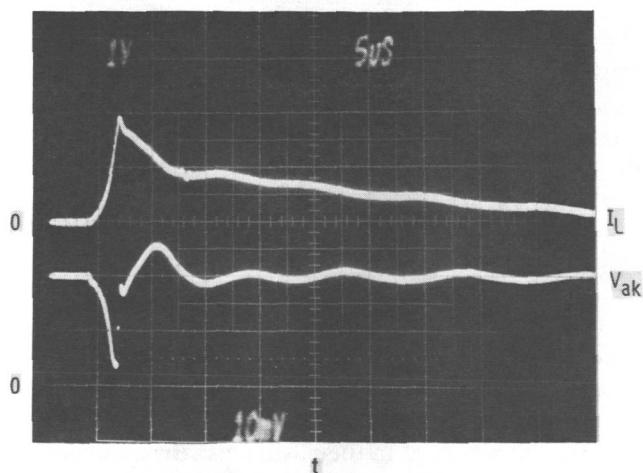Considering the packaging used for these RPC's, the specific weights achieved compare favorably with those

Figure 31.—Turnon of a GTO remote power controller into a short circuit. Load current I_L at 20 A/division; voltage across GTO V_{ak} at 100 V/division; load resistance, 0.85 Ω.
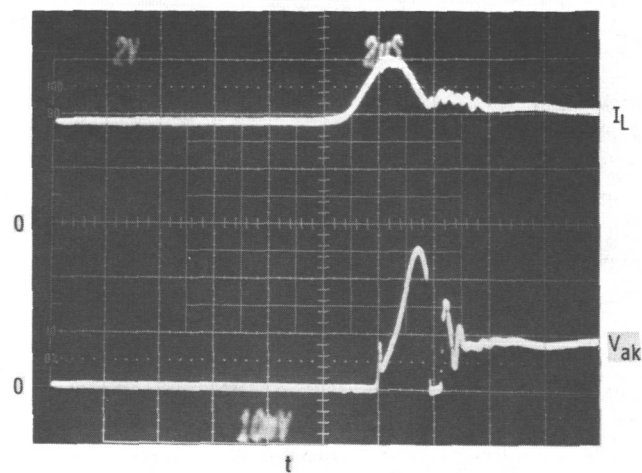

Figure 32.—Short circuit of GTO remote power controller from full load. Load current I_L at 20 A/division; voltage across GTO V_{ak} at 200 V/division; sweep, 2 μs/division.

TABLE VII.—GTO RPC THERMAL VACUUM
TEST DATA

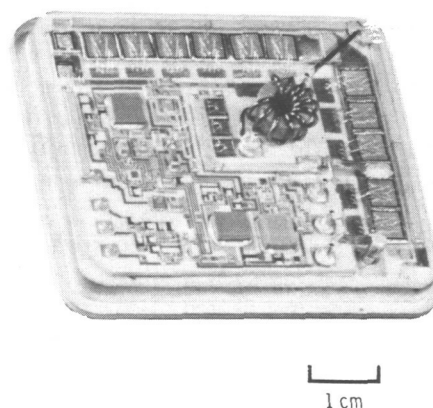|	Test		
	1	2	3
Mounting plate temperature, °C	80.4	56.9	27.9
Load current, A	34.99	34.97	34.94
Load voltage, V	129.6	129.5	129.5
GTO on-voltage, V	2.47	2.48	2.53
GTO heat sink temperature, °C	106	81.2	49.9
Baseplate temperature, °C	88.7	66.6	38.4

Figure 33.—Hybrid circuit implementation of 120-V dc, 30-A bipolar transistor remote power controller.

of previous units. As an example, the hybrid units developed for the space shuttle switch (32 V at 20 A) had a specific mass of 0.25 kg/kW. The MOSFET and GTO units built have specific masses of 0.17 to 0.21 and 0.04 to 0.18 kg/kW, respectively. Considerable improvement in these values is possible. Figure 33 shows a 120-V dc, 30-A solid-state RPC built in hybrid form on a ceramic substrate. It uses 11 power transistors, which can be seen along the upper and right edges of the substrate. At the 120-V level, an isolated drive stage was not needed. Adding this stage might at worst double the area required. It might also be possible to add a second layer. Either hybrid approach would work well for MOSFET RPC's because there are no bulky components and the internal power level is low. The only large add-on component would be the isolating power transformer. Even this could be miniaturized by using the techniques developed for commercial isolation amplifiers. With these techniques tenfold weight and volume reductions should be possible. This would make these units extremely attractive for all kinds of flight applications for both aircraft and space vehicles.

One major reason that this approach is practical is that chip MOSFET's can be easily mounted on a ceramic

substrate and can be electrically paralleled. GTO RPC's pose different problems. The GTO device can be considerably larger and bulky energy storage capacitors are required. Neither of these are insurmountable problems, but a hybrid GTO RPC will be somewhat larger than a similar MOSFET RPC.

Discussion and Conclusions

The technology of high-voltage, high-power, solid-state remote power controllers has been developed and demonstrated. Seven remote power controller units were developed that control a wide range of voltages and currents. Voltage-switching and power-handling capabilities are almost an order of magnitude greater than those of the state-of-art controllers used on the space shuttle. The basic circuits can, with slight modification, be used for remote power controllers limited only by the

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power rating of the switching device. The power controllers employed either power MOSFET's or GTO's as the main switching element. These devices are relatively new and it is expected that better ones, expanding the options available for space power systems, will continually be introduced.

With present devices some limitations must be observed to ensure reliable operation. Within the voltage, current, and forward-voltage drop limitations of the devices used, MOSFET RPC's can be used for all switching applications. Paralleling MOSFET's has been demonstrated to be very practical, so the only remaining limit is that on voltage. High-voltage MOSFET's are becoming available, but they have a high on-resistance and therefore many parallel devices must be used for high currents.

For higher voltage at high powers the GTO RPC is applicable. With 700-A, 2500-V devices available, single GTO switches to handle all near-term power requirements are possible. These applications include bus switching and load disconnects within the following limitations. Because of the uncontrollably fast turnon and turnoff times of GTO switches, they are difficult to protect from heavy overloads and short circuits. Thus they are not well suited for load switching in battery systems where large fault currents are possible. Capacitive input loads cannot be driven from such a source. In "softer" systems such as those powered by solar arrays, where there is a degree of current limiting, suitable GTO's can be found for most applications. It may be necessary to size the switch GTO on the basis of maximum possible surge current.

Another mitigating factor that would allow GTO RPC's to be used with sources that have high surge capability is the normal line inductance in medium-to-large systems. As transmission line lengths increase, the inductance can easily reach or exceed 10 μH, and, depending on line voltage and load, this may sufficiently limit current rise time.

If GTO switches can be tailored to the application, they can be used in many potential applications. It is important to realize that all of these RPC's have several orders of magnitude faster response than the electromechanical switches used in the past. This feature makes it possible to protect solid-state device loads, as cannot be done by using electromechanical breakers or even fast-acting fuses in many cases.

The RPC's demonstrated can be used in their present form for many applications. If sufficient quantities are required for a flight application, the next step should be to develop hybrid versions. With all components mounted on a ceramic substrate, weight and volume should decrease by a factor of 2 to 5 and further enlarge the area of application of remote power controllers.

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15. Supplementary Notes

16. Abstract

Two general types of remote power controller (RPC) that combine the functions of a circuit breaker and a switch were developed for use in direct-current (dc) aerospace systems. Power-switching devices used in these designs are the relatively new gate-turnoff thyristor (GTO) and power metal-oxide-semiconductor field-effect transistors (MOSFET). The various RPC's can switch dc voltages to 1200 V and currents to 100 A. Seven different units were constructed and subjected to comprehensive laboratory and thermal vacuum testing. Two of these were dual units that switch both positive and negative voltages simultaneously. The RPC's using MOSFET's have slow turnon and turnoff times to limit voltage spiking from high di/dt. The GTO's have much faster transition times. All RPC's have programmable overload tripout and microsecond tripout for large overloads. The basic circuits developed can be used to build switchgear limited only by the ratings of the switching device used.

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